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(54) SCALED AND ROTATED ALAMOUTI CODING

SKALIERTE UND ROTIERTE ALAMOUTI KODIERUNG

CODE D'ALAMOUTI CALIBRÉ ET ROTATIF

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EP 2 113 147 B1

Description

FIELD OF THE INVENTION

[0001] The present invention relates to an encoder and a corresponding encoding method for encoding incoming symbols of an incoming data stream into channel symbols of a channel data stream for transmission over a transmission channel.

[0002] Further, the present invention relates to a decoder being adapted for block by block decoding received channel symbols of a channel data stream, which have been encoded from incoming symbols of an incoming data stream, in particular by an encoder according to the present invention, and transmitted over a transmission channel.

[0003] The present invention relates also to a transmitter and a receiver, to a data signal encoded by an encoder according to the present invention and a computer program for implementing the encoding method in software.

BACKGROUND OF THE INVENTION

[0004] WO 99/14871 discloses a simple block coding arrangement in which symbols are transmitted over a plurality of transmit channels, in connection with coding that comprises only of simple arithmetic operations, such as negation and conjugation. The diversity created by the transmitter utilizes space diversity and either time or frequency diversity. Space diversity is effected by redundantly transmitting over a plurality of antennas, time diversity is effected by redundantly transmitting at different times, and frequency diversity is effected by redundantly transmitting at different frequencies. Illustratively, using two transmit antennas and a single receive antenna, one of the disclosed embodiments provides the same diversity gain as the maximal-ratio receiver combining (MRRC) scheme with one transmit antenna and two receive antennas.

[0005] The coding scheme disclosed in WO 99/14871 is known in the art as Alamouti coding scheme and has also been described in S.M. Alamouti "A simple transmit diversity technique for wireless communications", IEEE J. Sel. Areas. Comm. vol. 16, pp. 1451-1458, October 1998.

[0006] The Article "Path-Diversity for Phase Detection in Low-Cost Sensor Networks" of Althaus et al. discloses wireless sensor networks including some modifications of the Alamouti scheme for 2PSK.

[0007] The Article "Efficiency improvement for Alamouti codes" of Ling et al. discloses a method for increasing the spatial efficiency by transporting more information bits in each transmission block than the Alamouti code does.

[0008] The Article "Alamouti Transmit Diversity Scheme with a Simple Diagonal Weighing Matrix" of Li et al. discloses a method of Space Time Block Coding with a diagonal matrix, based on channel state information fed back to the transmitter.

SUMMARY OF THE INVENTION

[0009] It is an object of the present invention to provide a coding scheme, in particular an encoder, a corresponding encoding method and a decoder, having a better performance than the known Alamouti coding scheme.

[0010] The object is achieved according to the present invention by an encoder as claimed in claim 1, comprising:

- mapping means for block by block mapping incoming symbols onto pairs of channel symbols, a block comprising two incoming symbols, the mapping means being arranged for mapping the block onto two pairs of channel symbols such that said two pairs of channel symbols include scaled versions of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols, said scaled versions being obtained by applying a scaling function having having a scaling factor with an absolute value different from one and being piece-wise linear with at least two pieces, and
- output means for outputting said channel symbols.

[0011] According to an aspect of this disclosure, the mapping means are adapted for applying a scaling function for scaling said incoming symbols, said scaling function being dependent on the sign of one of the two incoming symbols, in particular of the second incoming symbol.

[0012] In another aspect, the mapping means are adapted for applying a scaling function for scaling said incoming symbols, said scaling function being dependent on the constellation of the two incoming symbols.

[0013] In still another aspect, the mapping means are adapted for applying a scaling function M(s) for scaling said incoming symbols s, said scaling function M being chosen such that M(M(s))=-s.

[0014] In still another aspect, the mapping means are adapted for applying a scaling function $M_2(s)$ for scaling said incoming symbols s, said scaling function $M_2(s)$ being chosen as $M_2(s)=2s-D_2(s)$, wherein D(s)=5b and b is the complex sign of s, defined as $b = \text{sign}(\text{Re}(s)) + j\text{sign}(\text{Im}(s))$.

[0015] In another aspect, the mapping means are adapted for applying a scaling function $M_3(s)$ for scaling said incoming symbols s, said scaling function $M_3(s)$ being chosen as $M_3(s)=3s+ D_3(s)$, wherein the sub-function $D_3(s)=X$ for large positive values of s, $D_3(s)=-X$ for large negative values of s and $D_3(s)=0$ for small positive or negative values of s and for s=0, X being an integer constant.

[0016] In another aspect, the output means are adapted for outputting said channel symbols to two transmission means, in particular two transmission antennas, for transmitting said channel symbols over said transmission channel.

[0017] The object is further achieved according to the present invention by a decoder as claimed in claim 9. In an aspect of this disclosure, the decoder comprises:

- selection means for selecting a pair of possible function values of incoming symbols or of possible function values of a scaled version of incoming symbols for decoding a current block of received channel symbols, wherein a block comprises a pair of received channel symbols,
- estimation means for determining an estimate of the selected pair of incoming symbols,
- calculating means for calculating the Euclidian distance between the received signal and said estimate,
- slicing means for slicing said estimate, and
- control means for repeating said steps with other pairs of possible function values of possible incoming symbols or of possible function values of a scaled version of incoming symbols until a predetermined stop condition is met or until a minimum Euclidian distance is found and for outputting said pair of sliced estimate of possible incoming symbols or of sliced estimate of said scaled version of possible incoming symbols resulting in the minimum Euclidian distance.

[0018] In an aspect, the subtraction means are adapted for applying as sub-function D_2 the function $D_2(s)=5b$, wherein b is the complex sign of s.

[0019] In another variant of this aspect, the deceoder further comprises decision means for deciding whether to decode a block of received channel symbols comprising two received channel symbols or a block of scaled versions of received channels symbols comprising two received channel symbols, which have been scaled by applying said scaling function, said decision being made based on channel estimates of the absolute values of the channel transfer functions of said transmission channel.

[0020] In still another aspect, the decision means are adapted for deciding to decode a block of received channel symbols comprising two received channel symbols if the condition

$$\text{symbols if the condition } |h_{12}|^2+|h_{22}|^2 \geq |h_{11}|^2+|h_{21}|^2$$

is met and to decode a block of scaled versions of received channels symbols comprising two received channel symbols, which have been scaled by applying said scaling function, if said condition is not met, where said parameters $h_{12}, h_{22}, h_{11}, h_{21}$ are the estimates of the channel transfer functions of said transmission channel.

[0021] The invention also relates to an encoding method, a transmitter, a receiver, an encoded data signal and a computer program as defined in further independent claims.

[0022] In addition to scaling incoming symbols and/or the complex conjugate of incoming symbols, it is preferred according to the present invention, in order to further improve the performance, that (the same and/or other) incoming symbols and/or the complex conjugate of (the same and/or other) incoming symbols are rotated by a rotation angle as defined in claim 2.

BRIEF DESCRIPTION OF THE DRAWINGS

[0023] The invention will now be explained in more detail with reference to the drawings in which

Fig.1 shows a block diagram of a general channel on which noise is added to a transmitted signal,

Fig.2 shows two mappings from x_{k1} to x_{k2}, showing on the right the scaled-repetition mapping, on the left the ordinary-repetition mapping,

Fig.3 shows a diagram illustrating the basic capacity C, the repetition capacity C_r the maximum transmission rates achievable with 4-PAM in the ordinary-repetition case Ia and the maximum rates achievable using scaled-repetition mapping Ib,

Fig.4 shows a model of a 2×2 MIMO channel,

Fig.5 shows a diagram illustrating the minimum modulus of the determinant for rotated and scaled Alamouti as a function of θ horizontally,

Fig.6 shows a diagram illustrating the message error rate for several R=4 space-time codes,

Fig.7 shows a diagram illustrating the message error rate for three rotated scaled Alamouti decoders (R = 4) (horizontally SNR),

Fig.8 shows a diagram illustrating the number of slicings for two rotated sclaed Alamouti decoders (R = 4) (horizontally SNR),

Fig.9 illustrates a further embodiment using scaled-repetition mapping $M_3(\cdot)$,

Fig.10 shows a diagram illustrating the message error rate for several R=6.34 space-time codes (horizontally SNR indB),

Fig.11 shows a diagram illustrating message error rate for three rotated scaled Alamouti decoders (R = 6.34) (horizontally SNR),

Fig.12 shows a diagram illustrating Number of slicings for two Rotated Scaled Alamouti decoders (R = 6.34) (horizontally SNR),

Fig.13 shows a block diagram of an embodiment of the present invention where two transmitter antennas and one receiver antenna are employed, and

Fig.14 shows a block diagram of an embodiment of the present invention an embodiment using two transmitter antennas and two receiver antennas are employed.

DETAILED DESCRIPTION OF EMBODIMENTS

[0024] First transmission over a single-input single-output (SISO) additive white Gaussian noise (AWGN) channel as shown in Fig.1 shall be considered, and scaled-repetition retransmission shall be introduced. It turns out that scaled-repetition improves upon ordinary-repetition retransmission.

[0025] First, some information theory shall be discussed. The real-valued output y_k for transmission $k = 1,2,\cdots, K$, see Fig.1, satisfies

$$y_k = x_k + n_k, \tag{1}$$

where x_k is the real-valued channel input for transmission k and n_k is a real-valued Gaussian noise sample with mean $E[N_k] = 0$, variance $E[N_k^2] = \sigma^2$, which is uncorrelated with all other noise samples. The transmitter power is limited, i.e. it is required that $E[X_k^2] \leq P$. It is well-known that an X which is Gaussian with mean 0 and variance P achieves capacity. This basic capacity (in bit/transm.) equals

$$C = \frac{1}{2}\log_2(1+\frac{P}{\sigma^2}). \tag{2}$$

[0026] When codewords are retransmitted (repeated), each symbol x_k from such a codeword $(x_1,x_2,\cdots, x_K)$ is actually transmitted and received twice, i.e. $x_{k1} = x_{k2} = x_k$, and

$$\begin{aligned} y_{k1} &= x_k + n_{k1}, \\ y_{k2} &= x_k + n_{k2}. \end{aligned} \tag{3}$$

[0027] An optimal receiver can form

$$z_k = \frac{y_{k1}+y_{k2}}{2} = x_k + \frac{n_{k1}+n_{k2}}{2}. \tag{4}$$

[0028] Now the variance of the noise variable $(N_{k1} + N_{k2})/2$ is $\sigma^2/2$. Therefore the repetition capacity for a single repetition in bit/transm. is

$$C_r = \frac{1}{4}\log_2(1+\frac{2P}{\sigma^2}). \tag{5}$$

[0029] Fig.3 shows the basic capacity C and repetition capacity C_r as a function of the signal-to-noise ratio SNR which is defined as

$$\mathrm{SNR} \stackrel{\Delta}{=} \frac{P}{\sigma^2}. \qquad (6)$$

[0030] It is easy to see that always $C_r \leq C$. For large SNR it can be written $C_r \approx C/2 + 1/4$, while for small SNR it is obtained $C_r \approx C$.

[0031] Next, ordinary and scaled repetition for 4-PAM (pulse-amplitude modulation) shall be discussed. When 4-PAM modulation is used, the channel inputs x_k assume values from $A_{4\text{-PAM}} = \{-3,-1,+1,+3\}$, each with probability 1/4. Ordinary repetition, see (3), leads to signal points $(x_1,x_2) = (x,x)$ for $x \in A_{4\text{-PAM}}$, see the left part of Fig.2. For this case the maximum transmission rate $I_a(X;Y_1,Y_2)$ is shown in Fig.3. Note that this maximum transmission rate is slightly smaller than the corresponding capacities C_r, mainly because uniform inputs are used instead of Gaussians.

[0032] Benelli's method (G. Benelli, "A new method for the integration of modulation and channel coding in an ARQ protocol," IEEE Trans. Commun., vol. COM-40, pp. 1594 - 1606, October 1992) can be used to improve upon ordinary-repetition retransmission, i.e. by modulating the retransmitted symbol differently. It could e.g. be taken

$$\begin{aligned} x_{k1} &= x_k, \\ x_{k2} &= M_2(x_k) \qquad \text{for} \qquad x_k \in A_{4\text{-PAM}}, \end{aligned} \qquad (7)$$

where $M_2(\alpha) = 2\alpha - 5$ if $\alpha > 0$ and $M_2(\alpha) = 2\alpha + 5$ for $\alpha < 0$. This method is called scaled repetition since a symbol is scaled by a factor (2 here) and then compensated (add -5 or +5) in order to obtain a symbol from $A_{4\text{-PAM}}$. This results in the signal points $(x,M_2(x))$ for $x \in A_{4\text{-PAM}}$, see Fig.2, right part. Also for the scaled-repetition case the maximum transmission rate $I_b(X;Y_1,Y_2)$ is shown in Fig.3. Note that this maximum transmission rate is only slightly smaller than the basic capacity C. Ordinary repetition is however definitively inferior to the basic transmission if the SNR is not very small.

[0033] Next, the demodulation complexity shall be discussed. Scaled repetition outperforms ordinary repetition, but also has a disadvantage. In an ordinary-repetition system the output $y_k = (y_{k1} + y_{k2})/2$ is simply sliced. In a system that uses scaled repetition it can only be sliced after having distinguished between two cases. More precisely note that

$$x_{k2} = M_2(x_k) = 2x_k - D_2(x_k), \qquad (8)$$

where $D_2(\alpha) = 5$ if $\alpha > 0$ and $D_2(\alpha) = -5$ if $\alpha < 0$. Now a slicer can be used for

$$\begin{aligned} y_{k1} + 2y_{k2} &= x_k + n_{k1} + 2(2x_k - D_2(x_k) + n_{k2}) \\ &= 5x_k - 2D_2(x_k) + n_{k1} + 2n_{k2}. \end{aligned} \qquad (9)$$

[0034] Assuming that $x_k \in \{-3,-1\}$ it is obtained that $D_2(x_k) = -5$ and this implies that a threshold shall be put at 0 to distinguish between -3 and -1. Similarly assuming that $x_k \in \{+1,+3\}$ it is obtained $D_2(x_k) = 5$ and it must be sliced $y_{k1} + 2y_{k2}$ again with a threshold at 0. Then the best overall candidate $\hat{x}_k$ is found by minimizing $(y_{k1} - \hat{x}_k)^2 + (y_{k2} - M_2(\hat{x}_k))^2$ over the two candidates.

[0035] Next, fundamental properties for the 2 x 2 MIMO channel shall be described and a model description shall be introduced. A 2 × 2 MIMO channel is shown in Fig.4. Both the transmitter T and the receiver R use two antennas. The output vector (y_{1k},y_{2k}) at transmission k relates to the corresponding input vector (x_{1k},x_{2k}) as given by

$$\begin{pmatrix} y_{1k} \\ y_{2k} \end{pmatrix} = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{pmatrix} \begin{pmatrix} x_{1k} \\ x_{2k} \end{pmatrix} + \begin{pmatrix} n_{1k} \\ n_{2k} \end{pmatrix} \qquad (10)$$

where (n_{1k},n_{2k}) is a pair of independent zero-mean circularly symmetric complex Gaussians, both having variance σ^2 (per two dimensions). Noise variable pairs in different transmissions are independent.

[0036] It is assumed that the four channel coefficients h_{11},h_{12},h_{21}, and h_{22} are independent zero-mean circularly symmetric complex Gaussians, each having variance 1 (per two dimensions). The channel coefficients are chosen prior to a block of K transmissions and remain constant over that block. The complex transmitted symbols (x_{k1}, x_{k2}) must satisfy a power constraint, i.e.

$$E[x_{k1}x_{k1}^{*} + x_{k1}x_{k2}^{*}] \leq P. \tag{11}$$

[0037] If the channel input variables are independent zero-mean circularly symmetric complex Gaussians both having variance $P/2$, then the resulting mutual information (called Telatar capacity here, see I.E. Telatar, "Capacity of multi-antenna Gaussian channels" European Trans. Telecommunications, vol. 10, pp. 585-595, 1999. (Originally published as AT&T Technical Memorandum, 1995)) is

$$C_{\text{Telatar}}(H) = \log_2 \det(I_2 + \frac{P/2}{\sigma^2} HH^{\dagger}), \tag{12}$$

where

$$H = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{pmatrix}, \tag{13}$$

i.e. the actual channel-coefficient matrix and I_2 the 2×2 identity matrix (here $H^{\dagger}$ denotes the Hermitian transpose of H. It involves both transposition and complex conjugation). Also in the 2×2 MIMO case the signal-to-noise ratio is defined as

$$\text{SNR} \stackrel{\Delta}{=} \frac{P}{\sigma^2}. \tag{14}$$

[0038] It can be shown (see e.g. H. Yao, "Efficient Signal, Code, and Receiver Designs for MIMO Communication Systems," Ph.D. thesis, M.I.T., June 2003, p. 36) that for fixed R and SNR large enough

$$\Pr\{C_{\text{Telatar}(H)} < R\} \approx \gamma \cdot \text{SNR}^{-4}, \tag{15}$$

for some constant γ.

[0039] The worst-case error-probabilities shall now be described. Consider M (one for each message) $K \times 2$ code-matrices $\underline{c}_1, \underline{c}_2, \cdots, \underline{c}_M$ resulting in a unit average energy code. Then Tarokh, Seshadri and Calderbank, "Space-Time Codes for High Data Rate Wireless Communication: Performance Criterion and Code Construction," IEEE Trans. Inform. Theory, Vol. 44, pp. 744- 765, March 1998, showed that for large SNR

$$\Pr\{\underline{c} \rightarrow \underline{c}'\} \approx \gamma'(\det((\underline{c}'-\underline{c})(\underline{c}'-\underline{c})^{\dagger})^{-2}\text{SNR}^{-4}. \tag{16}$$

for some γ' if the rank of the difference matrices $\underline{c}$-$\underline{c}'$ is 2, and it is transmitted $\underline{x} = \sqrt{P}\underline{c}$. If this holds for all difference matrices it is said that the diversity order is 4. Therefore it makes sense to maximize the minimum modulus of the determinant over all code-matrix differences.

[0040] S.M. Alamouti, "A simple transmit diversity technique for wireless communications," IEEE J. Sel. Areas. Comm. vol. 16, pp. 1451-1458, October 1998 proposed a modulation scheme (space-time code) for the 2×2 MIMO channel which allows for a very simple detector. Two complex symbols s_1 and s_2 are transmitted in the first transmission (an odd transmission) and in the second transmission (the next even transmission) these symbols are more or less repeated. More precisely

$$\begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} = \begin{pmatrix} s_1 & -s_2^{*} \\ s_2 & s_1^{*} \end{pmatrix}. \tag{17}$$

[0041] The received signal is now

$$\begin{pmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{pmatrix} = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{pmatrix} \begin{pmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{pmatrix} + \begin{pmatrix} n_{11} & n_{12} \\ n_{21} & n_{22} \end{pmatrix}. \tag{18}$$

[0042] Rewriting this results in

$$\begin{pmatrix} y_{11} \\ y_{21} \\ y_{12}^* \\ y_{22}^* \end{pmatrix} = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \\ h_{12}^* & -h_{11}^* \\ h_{22}^* & -h_{21}^* \end{pmatrix} \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} + \begin{pmatrix} n_{11} \\ n_{21} \\ n_{12}^* \\ n_{22}^* \end{pmatrix}. \tag{19}$$

or more compactly

$$\underline{y} = s_1 \underline{a} + s_2 \underline{b} + \underline{n}, \; (20)$$

with

$$\begin{aligned} \underline{y} &= (y_{11}, y_{21}, y_{12}^*, y_{22}^*)^T, \\ \underline{a} &= (h_{11}, h_{21}, h_{12}^*, h_{22}^*)^T, \\ \underline{b} &= (h_{12}, h_{22}, -h_{11}^*, -h_{21}^*)^T, \text{and} \\ \underline{n} &= (n_{11}, n_{21}, n_{12}^*, n_{22}^*)^T. \end{aligned} \tag{21}$$

[0043] Since $\underline{a}$ and $\underline{b}$ are orthogonal the symbol estimates $\hat{s}_1$ and $\hat{s}_2$ can be determined by simply slicing $(\underline{a}^\dagger \underline{y})/(\underline{a}^\dagger \underline{a})$ and $(\underline{y}^\dagger \underline{b})l(\underline{b}^\dagger \underline{b})$ respectively.

[0044] Another advantage of the Alamouti method is that the densities of $\underline{a}^\dagger \underline{a}$ and $\underline{b}^\dagger \underline{b}$ are (identical and) chi-square with 8 degrees of freedom. This results in a diversity order 4, i.e.

$$\Pr\{(\widehat{S_1, S_2}) \neq (S_1, S_2)\} \approx \gamma'' \cdot \mathrm{SNR}^{-4}, \tag{22}$$

for fixed rate and large enough SNR.

[0045] A disadvantage of the Alamouti method is that only two complex symbols are transmitted every two transmissions, but more-importantly that the symbols transmitted in the second transmission are more or less repetitions of the symbols in the first transmission. The above however suggests that ordinary repetition can be improved.

<u>The scaled Alamouti method</u>

[0046] Next, the scaled Alamouti method as proposed according to one embodiment of the present invention shall be described. Having seen above that scaled-repetition improves upon ordinary repetition in the SISO case, this concept can be used to improve upon the standard Alamouti scheme for MIMO transmision.

[0047] Instead of just repeating the symbols in the second transmission they are scaled (modulo the size of the signal constellation). More precisely, when s_1 and s_2 are elements of $A_{64-\mathrm{QAM}} \stackrel{\Delta}{=} \{a + jb \mid a \in A_{8-\mathrm{PAM}}, b \in A_{8-\mathrm{PAM}}\}$, it could be transmitted

$$\begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} = \begin{pmatrix} s_1 & -s_2^* \\ M(3s_2) & M(3s_1^*) \end{pmatrix} = \begin{pmatrix} s_1 & -s_2^* \\ 3s_2 & 3s_1^* \end{pmatrix} - \begin{pmatrix} 0 & 0 \\ D(3s_2) & D(3s_1^*) \end{pmatrix}. \tag{23}$$

where $M(\alpha) = \beta$ and $D(\alpha) = 16\gamma$ if $\beta \in A_{64\text{-QAM}}$ and there exists a complex number γ with integer components such that $\alpha = \beta + 16\gamma$.

[0048] In order to see what the perspective of the scaled-Alamouti method is, a Monte Carlo simulation has been carried out. Symbols from $A_{64\text{-QAM}}$ only have been considered. 2 × 2 MIMO-channels have been generated at random and it is computed for each sample channel the mutual information between channel input and output for the standard-Alamouti and the scaled-Alamouti case for s_1 and s_2 that are elements of $A_{64\text{-QAM}}$. Moreover the Telatar mutual information for each sample channel has been determined. Based on these mutual informations outage capacities have been computed. From this simulation it can be concluded that for rates $R < 5$ bit/channel use, the scaled-Alamouti method is not much worse than Telatar. Telatar assumes Gaussian input distributions and in the Alamouti cases a uniform $A_{64\text{-QAM}}$ distribution has been applied. This loss is small (roughly 0.1 bit/dimension) since the rate in bit per dimension is small (around 1 bit/dimension).

[0049] Furthermore, the rate of the scaled-Alamouti method is significantly larger than that of standard-Alamouti method (close to 1 bit/channel use for outage of 1 % at 14dB). Standard Alamouti is roughly 2dB worse than scaled Alamouti. In order to realize a certain rate at roughly the same outage probability the standard Alamouti method needs 2dB more signal power than scaled Alamouti, as shown in the table below:

	3 bit/chan. use	4 bit/chan. use	5 bit/chan. use
standard Alam.	8 dB 4 %	11 dB 6 %	14 dB 9 %
scaled Alam.	5 dB 10 %	8 dB 10 %	11 dB 9 %

[0050] In conclusion it can be said that from the perspective of outage-capacity scaled Alamouti is to be preferred over standard Alamouti. The disadvantage of scaled Alamouti is its larger decoding complexity. This will be discussed next.

[0051] In the scaled-Alamouti case the received vector is

$$\begin{pmatrix} y_{11} \\ y_{21} \\ y_{12}^* \\ y_{22}^* \end{pmatrix} = \begin{pmatrix} h_{11} & 3h_{12} \\ h_{21} & 3h_{22} \\ 3h_{12}^* & -h_{11}^* \\ 3h_{22}^* & -h_{21}^* \end{pmatrix} \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} - \begin{pmatrix} 0 \\ 0 \\ h_{12}^* \\ h_{22}^* \end{pmatrix} D(3s_1) - \begin{pmatrix} h_{12} \\ h_{22} \\ 0 \\ 0 \end{pmatrix} D(s_2) + \begin{pmatrix} n_{11} \\ n_{21} \\ n_{12}^* \\ n_{22}^* \end{pmatrix}. \tag{24}$$

[0052] This can be written as

$$\underline{y} = s_1\underline{a} + s_2\underline{b} - D(3s_1)\underline{c} - D(3s_2)\underline{d} + \underline{n},$$

with $\underline{y} = (y_{11}, y_{21}, y_{12}^*, y_{22}^*)^T$, $\underline{a} = (h_{11}, h_{21}, 3h_{12}^*, 3h_{22}^*)^T$, $\underline{b} = (3h_{12}, 3h_{22}, -h_{11}^*, -h_{21}^*)^T$, $\underline{c} = (0, 0, h_{12}^*, h_{22}^*)^T$, $\underline{d} = (h_{12}, h_{22}, 0, 0)^T$, and $\underline{n} = (n_{11}, n_{21}, n_{12}^*, n_{22}^*)^T$.

[0053] It is important to note that, just like in the standard-Alamouti case,

$$\underline{a}^{\dagger}\underline{b}=0,$$

hence $\underline{a}$ and $\underline{b}$ are orthogonal.

[0054] An optimal detector determines
$$\arg\min_{s_1,s_2} |\underline{y}-s_1\underline{a}-s_2\underline{b}+D(3s_1)\underline{c}+D(3s_2)|^2 .$$

[0055] For message-error-rates less than 0.01, scaled Alamouti is roughly 3dB better than standard Alamouti. For smaller message-error rates the difference becomes smaller.

[0056] Next, the decoding method is explained for the scaled Alamouti method.

[0057] The first proposed algorithm now checks all possible offset combinations $D(3s_1)$ and $D(3s_2)$. Note that both offsets can assume values $d = d' + jd''$ with $d' \in \{-16,0,+16\}$ and $d'' \in \{-16,0,+16\}$. This implies that 81 offset combinations are possible and have to be checked. Checking a combination requires offset correction i.e. computing

$$\underline{z}=\underline{y}+D(3s_1)\underline{c}+D(3s_2)\underline{d}.$$

[0058] Then both $\underline{a}^{\dagger}\underline{z}/\underline{a}^{\dagger}\underline{a}$ and $\underline{b}^{\dagger}\underline{z}/\underline{b}^{\dagger}\underline{b}$ are sliced in a restricted way, i.e. only those $\hat{s}_1$ and $\hat{s}_2$ are considered that have the assumed $D(3\hat{s}_1)$ and $D(3\hat{s}_2)$. This results in a distance metric for all of the 81 alternatives from which the best one is chosen.

[0059] A first improvement follows from the fact that it is not needed to slice if the distance of the received vector to a signal point will be too large anyhow. A lower bound for this distance can be obtained by considering

$$\underline{z}^{\perp}=\underline{z}-\frac{\underline{a}^{\dagger}\underline{z}}{\underline{a}^{\dagger}\underline{a}}\underline{a}-\frac{\underline{b}^{\dagger}\underline{z}}{\underline{b}^{\dagger}\underline{z}}\underline{b},$$

i.e. the part of $\underline{z}$ perpendicular to both $\underline{a}$ and $\underline{b}$. Then

$$\begin{aligned} |\underline{y}-s_1\underline{a}-s_2\underline{b}+D(3s_1)\underline{c}+D(3s_2)|^2 &= |\underline{z}-s_1\underline{a}-s_2\underline{b}|^2 \\ &= |\underline{z}^{\perp}|^2+|\underline{z}-\underline{z}^{\perp}-s_1\underline{a}-s_2\underline{b}|^2 \\ &\geq |\underline{z}^{\perp}|^2, \end{aligned} \tag{25}$$

and if $|\underline{z}^{\perp}|^2$ is larger than (or equal to) the smallest squared distance observed for an offset-combination before, slicing is unnecessary.

[0060] Note furthermore that

$$\underline{z}^{\perp}=\underline{y}^{\perp}+D(3s_1)\underline{c}^{\perp}+D(3s_2)\underline{d}^{\perp}, \tag{26}$$

with

$$\begin{aligned} \underline{y}^{\perp} &= \underline{y}-\frac{\underline{a}^{\dagger}\underline{y}}{\underline{a}^{\dagger}\underline{a}}\underline{a}-\frac{\underline{b}^{\dagger}\underline{y}}{\underline{b}^{\dagger}\underline{b}}\underline{b} \\ \underline{c}^{\perp} &= \underline{c}-\frac{\underline{a}^{\dagger}\underline{c}}{\underline{a}^{\dagger}\underline{a}}\underline{a}-\frac{\underline{b}^{\dagger}\underline{c}}{\underline{b}^{\dagger}\underline{b}}\underline{b} \\ \underline{d}^{\perp} &= \underline{c}-\frac{\underline{a}^{\dagger}\underline{d}}{\underline{a}^{\dagger}\underline{a}}\underline{a}-\frac{\underline{b}^{\dagger}\underline{d}}{\underline{b}^{\dagger}\underline{b}}\underline{b}, \end{aligned} \tag{27}$$

hence $\underline{z}^{\perp}$ is a simple linear combination of $\underline{y}^{\perp}$ and $\underline{c}^{\perp}$, and $\underline{d}^{\perp}$ and can be computed easily if $D(3s_1)$ and $D(3s_2)$ vary.

[0061] In the second method the amount of work is smaller if the first checked offset combination would already yield a small squared distance. This can be achieved by making an estimate of both $D(3s_1)$ and $D(3s_2)$ based on $\underline{y}$. Therefore consider

$$\begin{aligned}\begin{pmatrix} y_{11} \\ y_{21} \end{pmatrix} &= \begin{pmatrix} h_{11} & 3h_{12} \\ h_{21} & 3h_{22} \end{pmatrix}\begin{pmatrix} s_1 \\ s_2 \end{pmatrix} - \begin{pmatrix} h_{12} \\ h_{22} \end{pmatrix} D(3s_2) + \begin{pmatrix} n_{11} \\ n_{21} \end{pmatrix} \\ &= \begin{pmatrix} h_{11} \\ h_{21} \end{pmatrix} s_1 + \begin{pmatrix} h_{12} \\ h_{22} \end{pmatrix} M(3s_2) + \begin{pmatrix} n_{11} \\ n_{21} \end{pmatrix}\end{aligned} \tag{28}$$

and assume that the second and third term are noise terms. Then it can be computed $\frac{\underline{h}(1)^{\dagger}\underline{y}(1)}{\underline{h}(1)^{\dagger}\underline{h}(1)}$ to get an estimate of s_1 and subsequently of $D(3s_1)$. Here

$$\begin{aligned}\underline{y}(1) &= (y_{11}, y_{21})^T \\ \underline{h}(1) &= (h_{11}, h_{21})^T.\end{aligned} \tag{29}$$

[0062] Similarly consider

$$\begin{aligned}\begin{pmatrix} y_{12}^* \\ y_{22}^* \end{pmatrix} &= \begin{pmatrix} 3h_{12}^* & -h_{11}^* \\ 3h_{22}^* & -h_{21}^* \end{pmatrix}\begin{pmatrix} s_1 \\ s_2 \end{pmatrix} - \begin{pmatrix} h_{12}^* \\ h_{22}^* \end{pmatrix} D(3s_1) + \begin{pmatrix} n_{12}^* \\ n_{22}^* \end{pmatrix} \\ &= \begin{pmatrix} -h_{11}^* \\ -h_{21}^* \end{pmatrix} s_2 + \begin{pmatrix} h_{12}^* \\ h_{22}^* \end{pmatrix} M(3s_1) + \begin{pmatrix} n_{12}^* \\ n_{22}^* \end{pmatrix}.\end{aligned}$$

[0063] It can now be computed

$$\frac{\underline{h}(2)^{\dagger}\underline{h}(2)}{\underline{h}(2)^{\dagger}\underline{h}(2)} \tag{30}$$

to get an estimate of s_2 and subsequently of $D(3s_2)$. Here

$$\begin{aligned}\underline{y}(2) &= (y_{12}^*, y_{22}^*)^T \\ \underline{h}(2) &= (-h_{11}^*, -h_{21}^*)^T.\end{aligned} \tag{31}$$

[0064] The initial guesses for $D(3s_1)$ and $D(3s_2)$ are now used to obtain an initial estimate of the signal pair (s_1,s_2). The associated squared distance is used in the remaining part of the decoding procedure.

The rotated and scaled Alamouti method

[0065] Next, the rotated and scaled Alamouti method as proposed according to a further embodiment of the present invention shall be described. Having seen above that scaled-repetition improves upon ordinary repetition in the SISO case, this concept is used to improve upon the standard Alamouti scheme for MIMO transmission. Instead of just repeating the symbols in the second transmission they are scaled. More precisely, when s_1 and s_2 are elements of $A_{16\text{-QAM}} \overset{\Delta}{=} \{a + jb \mid a \in A_{4\text{-PAM}}, b \in A_{4\text{-PAM}}\}$, for some value of θ the signals

$$\begin{aligned}\begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} &= \begin{pmatrix} s_1 \cdot \exp(j\theta) & -s_2^* \\ M_2(s_2) & M_2(s_1^*) \end{pmatrix} \\ &= \begin{pmatrix} s_1 \cdot \exp(j\theta) & -s_2^* \\ 2s_2 & 2s_1^* \end{pmatrix} - \begin{pmatrix} 0 & 0 \\ D_2(s_2) & D_2(s_1^*) \end{pmatrix},\end{aligned} \tag{32}$$

are transmited where $M_2(\alpha) = 2\alpha - D_2(\alpha)$ with $D_2(\alpha) = 5\beta$ when β is the complex sign of α, defined as $\beta = \mathrm{sign}(\mathrm{Re}(\alpha)) + j\mathrm{sign}(\mathrm{Im}(\alpha))$.

[0066] A first question is to determine a good value for θ. Therefore for $0 \leq \theta \leq \pi/2$ the minimum modulus of the determinant mindet(θ)

$$\mathrm{mindet}(\theta) = \min_{(s_1,s_2),(s_1',s_2')} |\det(X(s_1,s_2,\theta) - X(s_1',s_2',\theta))|, \tag{33}$$

is determined where

$$X = \begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} \tag{34}$$

is the code matrix. The minimum modulus of the determinant as a function of θ can be found in Fig.5. The maximum value of the minimum determinant (i.e. 7.613) occurs for

$$\theta_{\text{opt.}} = 1.028. \tag{35}$$

[0067] This value for θ will be used in the steps explained in the following.

[0068] Next, the hard-decision performance shall be discussed. The message-error-rate for several R = 4 space-time codes has been compared in Fig.6. By message-error-rate the probability $\Pr\{\hat{X} \neq X\}$ is meant. Note that for each "test" a new message (8-bit) and a new channel matrix have been prepared. The decoder is optimal for all codes, it performs *ML* -decoding (exhaustive search). The methods that have been considered are:

- Uncoded (B): It is transmitted

$$X = \begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix}, \tag{36}$$

  where x_{11}, x_{12}, x_{21}, and x_{22} are symbols from $A_{4\text{-QAM}}$.
- Alamouti (C): see (17), where s_1 and s_2 are symbols from $A_{16\text{-QAM}}$.
- Tilted QAM (E): Proposed by H. Yao and G.W. Wornell, "Achieving the full MIMO diversity-multiplexing frontier with rotation-based space-time codes," in Proc. Allerton Conf. Commun. Control, and Comput., Monticello, IL, Oct. 2003. Let s_a, s_b, s_c, and s_d symbols from $A_{4\text{-QAM}}$. Then it is transmitted

$$\begin{aligned}\begin{pmatrix} x_{11} \\ x_{22} \end{pmatrix} &= \begin{pmatrix} \cos(\theta_1) & -\sin(\theta_1) \\ \sin(\theta_1) & \cos(\theta_1) \end{pmatrix} \begin{pmatrix} s_a \\ s_b \end{pmatrix}, \\ \begin{pmatrix} x_{21} \\ x_{12} \end{pmatrix} &= \begin{pmatrix} \cos(\theta_2) & -\sin(\theta_2) \\ \sin(\theta_2) & \cos(\theta_2) \end{pmatrix} \begin{pmatrix} s_c \\ s_d \end{pmatrix},\end{aligned} \tag{37}$$

  for

$$\begin{aligned}\theta_1 &= \frac{1}{2}\arctan(\frac{1}{2}),\\ \theta_2 &= \frac{1}{2}\arctan(2).\end{aligned} \tag{38}$$

- Rotated and scaled Alamouti (D): see (32) for $\theta = 1.028$, and with s_1 and s_2 from $A_{16\text{-QAM}}$.
- Golden code (F): Proposed by J.-C. Belfiore, G. Rekaya, E. Viterbo, "The golden code: A 2x 2 full-rate space-time code with nonvanishin determinants," IEEE Trans. Inform. Theory, vol. IT-51, No. 4, pp. 1432 - 1436, April 2005. Now

$$X = \frac{1}{\sqrt{5}}\begin{pmatrix} \alpha(z_1 + z_2\theta) & \alpha(z_3 + z_4\theta) \\ j \cdot \overline{\alpha}(z_3 + z_4\overline{\theta}) & \overline{\alpha}(z_1 + z_2\overline{\theta}) \end{pmatrix}, \tag{39}$$

with $\theta = \frac{1+\sqrt{5}}{2}$, $\overline{\theta} = \frac{1-\sqrt{5}}{2}$, $\alpha = 1 + j - j\theta$, and $\overline{\alpha} = 1 + j - j\overline{\theta}$ and where z_1, z_2, z_3, and z_4 are $A_{4\text{-QAM}}$ -symbols.
- Telatar (A): This is the probability that the Telatar capacity of the channel is smaller than 4.

[0069] Clearly it follows from Fig.6 that the Golden code shows the best result. However rotated and scaled Alamouti is only slightly worse, roughly 0.2 dB. Important is that Alamouti coding is roughly 2dB worse than the Golden code.

[0070] The decoding complexity shall be discussed next. Clearly the Golden code is better than rotated and scaled Alamouti. However the Golden code in principle requires the decoder to check all 256 alternative codewords. Here the complexity and performance of a suboptimal rotated and scaled Alamouti decoder will be investigated. Denote

$$\Theta = \exp(j\theta_{\text{opt.}}).$$

[0071] In the rotated and scaled Alamouti case the received vector is

$$\begin{aligned}\begin{pmatrix} y_{11} \\ y_{21} \\ y_{12}^* \\ y_{22}^* \end{pmatrix} &= \begin{pmatrix} h_{11}\Theta & 2h_{12} \\ h_{21}\Theta & 2h_{22} \\ 2h_{12}^* & -h_{11}^* \\ 2h_{22}^* & -h_{21}^* \end{pmatrix}\begin{pmatrix} s_1 \\ s_2 \end{pmatrix} - \begin{pmatrix} 0 \\ 0 \\ h_{12}^* \\ h_{22}^* \end{pmatrix} D_2(s_1) \\ &\quad - \begin{pmatrix} h_{12} \\ h_{22} \\ 0 \\ 0 \end{pmatrix} D_2(s_2) + \begin{pmatrix} n_{11} \\ n_{21} \\ n_{12}^* \\ n_{22}^* \end{pmatrix}.\end{aligned} \tag{40}$$

[0072] This can be written as

$$\underline{y} = s_1\underline{a} + s_2\underline{b} - D_2(s_1)\underline{c} - D_2(s_2)\underline{d} + \underline{n},$$

with $\underline{y} = (y_{11}, y_{21}, y_{12}^*, y_{22}^*)^T$, $\underline{a} = (h_{11}\Theta, h_{21}\Theta, 2h_{12}^*, 2h_{22}^*)^T$, $\underline{b} = (2h_{12}, 2h_{22}, -h_{11}^*, -h_{21}^*)^T$, $\underline{c} = (0, 0, h_{12}^*, h_{22}^*)^T$, $\underline{d} = (h_{12}, h_{22}, 0, 0)^T$, and $\underline{n} = (n_{11}, n_{21}, n_{12}^*, n_{22}^*)^T$.

[0073] For the angle ϕ between $\underline{a}$ and $\underline{b}$ it can be written

$$\cos\phi = \frac{\Re[2(\Theta-1)(h_{11}h_{12}^* + h_{21}h_{22}^*)]}{|h_{11}|^2 + |h_{21}|^2 + 4|h_{12}|^2 + 4|h_{22}|^2}. \tag{41}$$

[0074] Instead of decoding (s_1,s_2) it is also possible to decode $(t_1,t_2) = (M_2(s_1),(M_2(s_2))$ which is equivalent to (s_1,s_2). Therefore (32) is rewritten to obtain

$$\begin{aligned}\begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} &= \begin{pmatrix} -M_2(t_1)\Theta & M_2(t_2^*) \\ t_2 & t_1^* \end{pmatrix} \\ &= \begin{pmatrix} -2t_1\Theta & 2t_2^* \\ t_2 & t_1^* \end{pmatrix} - \begin{pmatrix} -D_2(t_1)\Theta & D_2(t_2^*) \\ 0 & 0 \end{pmatrix},\end{aligned} \tag{42}$$

since $t = M_2(s)$ implies that $s = -M_2(t)$.

$$\begin{aligned}\begin{pmatrix} y_{11} \\ y_{21} \\ y_{12}^* \\ y_{22}^* \end{pmatrix} &= \begin{pmatrix} -2h_{11}\Theta & h_{12} \\ -2h_{21}\Theta & h_{22} \\ h_{12}^* & 2h_{11}^* \\ h_{22}^* & 2h_{21}^* \end{pmatrix} \begin{pmatrix} t_1 \\ t_2 \end{pmatrix} - \begin{pmatrix} -h_{11}\Theta \\ -h_{21}\Theta \\ 0 \\ 0 \end{pmatrix} D_2(t_1) \\ &\quad - \begin{pmatrix} 0 \\ 0 \\ h_{11}^* \\ h_{21}^* \end{pmatrix} D_2(t_2) + \begin{pmatrix} n_{11} \\ n_{21} \\ n_{12}^* \\ n_{22}^* \end{pmatrix}.\end{aligned} \tag{43}$$

[0075] This can be written as

$$\underline{y} = t_1\underline{a}' + t_2\underline{b}' - D_2(t_1)\underline{c}' - D_2(t_2)\underline{d}' + \underline{n},$$

with $\underline{a}' = (-2h_{11}\Theta, -2h_{21}\Theta, h_{12}^*, h_{22}^*)^T$, $\underline{b}' = (h_{12}, h_{22}, 2h_{11}^*, 2h_{21}^*)^T$, $\underline{c}' = (-h_{11}\Theta, -h_{21}\Theta, 0, 0)^T$, and $\underline{d}' = (0,0,h_{11}^*,h_{21}^*)^T$, and for the angle ϕ' between $\underline{a}'$ and $\underline{b}'$ the can be written as

$$\cos\phi' = \frac{\Re[2(\Theta-1)(h_{11}h_{12}^* + h_{21}h_{22}^*)]}{4|h_{11}|^2 + 4|h_{21}|^2 + |h_{12}|^2 + |h_{22}|^2}. \tag{44}$$

[0076] It now follows from the inequality $2r_1r_2 \le r_1^2 + r_2^2$ (where r_1 and r_2 are reals), that

$$\begin{aligned}\cos\phi &\le |\Theta-1| \cdot \frac{|h_{11}|^2 + |h_{12}|^2 + |h_{21}|^2 + |h_{22}|^2}{|h_{11}|^2 + |h_{21}|^2 + 4|h_{12}|^2 + 4|h_{22}|^2}, \\ \cos\phi' &\le |\Theta-1| \cdot \frac{|h_{11}|^2 + |h_{12}|^2 + |h_{21}|^2 + |h_{22}|^2}{4|h_{11}|^2 + 4|h_{21}|^2 + |h_{12}|^2 + |h_{22}|^2}.\end{aligned} \tag{45}$$

If

$$|h_{12}|^2 + |h_{22}|^2 \geq |h_{11}|^2 + |h_{21}|^2, \tag{46}$$

then

$$\cos\phi \leq \frac{2|\Theta - 1|}{5} = 0.393, \tag{47}$$

else

$$\cos\phi' \leq \frac{2|\Theta - 1|}{5} = 0.393. \ (48)$$

[0077] Therefore it makes sense to decode (s_1,s_2) when (46) holds and (t_1,t_2) when (46) does not hold. Using zero-forcing to decode, the noise enhancement is then at most 1/(1- 0.3932) = 1.183 which is 0.729dB. It will be seen below that noise enhancement turns out to be un-noticeable in practise.

[0078] The decoding procedure is straightforward. Focus on the case where it is decoded (s_1,s_2) for a moment. For all 16 alternatives of (($D_2(s_1)$,$D_2(s_2)$)) the vector

$$\underline{z} = \underline{y} + D_2(s_1)\underline{c} + D_2(s_2)\underline{d} = s_1\underline{a} + s_2\underline{b} + \underline{n} \tag{49}$$

and is determined. Then the sufficient statistic

$$\begin{pmatrix} \underline{a}^\dagger \underline{z} \\ \underline{b}^\dagger \underline{z} \end{pmatrix} = \begin{pmatrix} \underline{a}^\dagger \underline{a} & \underline{a}^\dagger \underline{b} \\ \underline{b}^\dagger \underline{a} & \underline{b}^\dagger \underline{b} \end{pmatrix} \begin{pmatrix} s_1 \\ s_2 \end{pmatrix} + \begin{pmatrix} \underline{a}^\dagger \underline{n} \\ \underline{b}^\dagger \underline{n} \end{pmatrix}. \tag{50}$$

[0079] Next the inverted matrix

$$M = \begin{pmatrix} \underline{b}^\dagger \underline{b} & -\underline{a}^\dagger \underline{b} \\ -\underline{b}^\dagger \underline{a} & \underline{a}^\dagger \underline{a} \end{pmatrix} / D \tag{51}$$

is used where $D = (\underline{a}^\dagger\underline{a})(\underline{b}^\dagger\underline{b}) - (\underline{b}^\dagger\underline{a})(\underline{a}^\dagger b)$ to obtain

$$\begin{pmatrix} \tilde{s}_1 \\ \tilde{s}_2 \end{pmatrix} = M \begin{pmatrix} \underline{a}^\dagger \underline{z} \\ \underline{b}^\dagger \underline{z} \end{pmatrix}. \tag{52}$$

[0080] Next both $\tilde{s}_1$ and $\tilde{s}_2$ are sliced under the restriction that only alternatives that match the assumed values $D_2(s_1)$ and $D_2(s_2)$ are possible outcomes. This is done for all 16 alternatives ($D_2(s_1)$,$D_2(s_2)$). The best result in terms of Euclidean distance is now chosen.

[0081] In considering all alternatives (($D_2(s_1)$,$D_2(s_2)$)) it is only required to slice when the length of

$$\underline{z} - \tilde{s}_1\underline{a} - \tilde{s}_2\underline{b} \tag{53}$$

is smaller than the closest distance that has been observed so far. This reduces the number of slicing steps. This approach shall be called method 1.

[0082] The number of slicing steps can even be further decreased if decoding is started with the most promising alternative $((D_2(s_1),D_2(s_2))$. This approach is called method 2. Therefore it is noted that the "direct" s_1 -signal-component in X is

$$\begin{pmatrix} s_1\Theta & 0 \\ 0 & -s_1^*/2 \end{pmatrix}. \tag{54}$$

[0083] Therefore $(\underline{e}_1^\dagger \underline{y})/(\underline{e}_1^\dagger \underline{e}_1)$ can be sliced in order to find a good guess for $D_2(s_1)$. Similarly $(\underline{e}_2^\dagger \underline{y})/(\underline{e}_2^\dagger \underline{e}_2)$ is sliced to find a good first guess for $D_2(s_2)$. Here

$$\begin{aligned} \underline{e}_1 &= (h_{11}\Theta, h_{21}\Theta, -h_{12}^*/2, -h_{22}^*/2)^T, \\ \underline{e}_2 &= (-h_{12}/2, -h_{22}/2, -h_{11}^*, -h_{21}^*)^T. \end{aligned} \tag{55}$$

[0084] Then the other 15 alternatives are considered and only sliced if necessary. Note that similar methods apply if (t_1,t_2) shall be decoded.

[0085] Simulations have been carried out, first to find out what the degradation of the suboptimal decoders according to method 1 and method 2 is relative to ML-decoding. The result is shown in Fig.7. The conclusion is that the suboptimal decoders do not demonstrate a performance degradation.

[0086] The average number of slicings for both method 1 and method 2 have also been considered. This is shown in Fig.8. It can be observed that method 1 leads to roughly 7 slicings on average (as opposed to 16). Method 2 further decreases the average number of slicing to roughly 3.5.

[0087] The rotated and scaled Alamouti can also be based on 9 -PAM as explained in the following. The rate of the code that was considered in the previous sections is 4 bit per channel use. To increase this rate it turns out that it is important to start from a PAM constellation with a square number of points. Therefore the next constellation is 9 -PAM. The mapping that has been considered is $M_3(\cdot)$ which is defined as updated

$$M_3(x) = \begin{cases} 3x+20 & \text{for} \quad x \in \{-8,-6,-4\} \\ 3x & \text{for} \quad x \in \{-2,0,+2\} \\ 3x-20 & \text{for} \quad x \in \{+4,+6,+8\}. \end{cases} \tag{56}$$

It is important is that this mapping satisfies $M_3(M_3(x)) = -x$, see Fig.9. There are three intervals each containing three points.

[0088] More generally,

$$M_3(x) = 3x - D_3(x)$$

where it holds for the subfunction $D_3(x) = \begin{cases} +20 & \text{for} \quad x \in \{-8,-6,-4\} \\ 0 & \text{for} \quad x \in \{-2,0,+2\} \\ -20 & \text{for} \quad x \in \{+4,+6,+8\} \end{cases}$ in the above example.

[0089] Now, a rotated and scaled Alamouti method based on this mapping and operating on symbols s_1 and s_2 from $A_{81-\mathrm{QAM}} \stackrel{\Delta}{=} \{a+jb \mid a \in A_{9-\mathrm{PAM}}, b \in A_{9-\mathrm{PAM}}\}$ can be designed. The optimal value of θ = 1.308. This method can now again be compared with corresponding uncoded, Alamouti, Tilted-QAM, and Golden code methods. The results are shown in Fig.10. It is clear again that the Golden code has the best performance. Rotated and scaled Alamouti is again worse, roughly 0.5 dB, but Alamouti is roughly 4dB worse than the Golden code.

[0090] Again, simulations have been made to find out what the degradation of the suboptimal decoders according to method 1 and method 2 is relative to ML-decoding. The result is shown in Fig.11. Again the conclusion is that the suboptimal decoders do not demonstrate a performance degradation. The number of slicings for both method 1 and method 2 are shown in Fig.12. It can be observed that method 1 leads to roughly 21 slicings on average (as opposed to 81). Method 2 further decreases the average number of slicing to roughly 10. Note that exhaustive search here requires checking $81^2 = 6561$ codewords.

[0091] The conclusion is that the rotated and scaled Alamouti method as proposed according to the present invention has a hard-decision performance which is only slightly worse than that of the Golden code, but can be decoded with an acceptable complexity.

[0092] Figs. 13 and 14 show block diagrams of two particular embodiments of transmitters and receivers according to the present invention. The general function and working of these transmitters and receivers has been described in WO 99/14871 (to which explanations reference is explicitly made here) and shall thus not be explained here in all details. These transmitters and receivers are adapted to such that the elements thereof can carry out the steps of the methods of the present invention as described above.

[0093] Fig.13 shows a block diagram of an embodiment where two (generally k) transmitter antennas (providing space diversity) employing multiple time intervals and one receiver antenna are employed. Specifically, transmitter 10 illustratively comprises antennas 11 and 12, and it handles incoming data in blocks of 2 (generally k) symbols, where k is the number of transmitter antennas. Each block takes 2 (generally k) symbol intervals to transmit. Also illustratively, the FIG.1 arrangement includes a receiver 20 that comprises a single antenna 21.

[0094] At any given time, a signal sent by a transmitter antenna experiences interference effects of the traversed channel, which consists of the transmit chain, the air-link, and the receive chain. The channel may be modeled by a complex multiplicative distortion factor composed of a magnitude response and a phase response. Noise from interference and other sources is added at the two received signals, i.e. the resulting baseband signal received at any time and outputted by reception and amplification section 25 includes such noise in addition to the transmitted signals.

[0095] The received signal is applied to channel estimator 22, which provides signals representing the channel characteristics or, rather, the best estimates thereof.

[0096] Those signals are applied to combiner 23 and to maximum likelihood detector 24.

[0097] The estimates developed by channel estimator 22 can be obtained by sending a known training signal that channel estimator 22 recovers, and based on the recovered signal the channel estimates are computed. This is a well known approach.

[0098] Combiner 23 receives the signal in the first time interval, buffers it, receives the signal in the next time interval, and combines the two received signals to develop estimates of the transmitted signals.

[0099] These signal estimates are sent to maximum likelihood detector 24, which develops the transmitted signals with the aid channel estimates from estimator 22.

[0100] Fig.14 presents an embodiment where two transmit antennas 31, 32 and two receive antennas 51, 52 are used. The signal received by antenna 51 is applied to channel estimator 53 and to combiner 55, and the signal received by antenna 52 is applied to channel estimator 54 and to combiner 55. Estimates of the channel transfer functions from the transmit antennas 31, 32 to the receive antenna 51 are applied by channel estimator 53 to combiner 55 and to maximum likelihood detector 56. Similarly, estimates of the channel transfer functions from the transmit antennas 31, 32 to the receive antenna 52 are applied by channel estimator 54 to combiner 55 and to maximum likelihood detector 56. There, the transmitted signals are recovered.

[0101] The invention has been explained above with reference to embodiments referring a 2x2 MIMO system. However, in the most general sense, the idea underlying the present invention refers to the building of a block code mapping a k-symbol vector into an n-symbol vector, in which the n symbols are scaled (and, more preferably, rotated) versions of the k symbols or of the complex conjugate of the k symbols with at least one scaling function being piece-wise linear with at least two pieces.

[0102] The code is then built for a MIMO system with k transmit antennas. Hence, the "pairs" as, for instance, mentioned in the claims, might as well be triplets or tuples with an appropriately adapted mapping, and the invention could be applied in any MIMO system. For instance, three complex symbols (a triplet) could be mapped onto 2, 3, or more symbol-triplet using scaling (and rotation). Further, the embodiment of the invention about scaled repetition is preferably used in ARQ SISO systems, in which the retransmission is a scaled version of the original transmitted symbol.

[0103] Still further, instead of one MIMO transmitter, also a distributed transmitter could use the present invention of a rotated and scaled Alamouti method. It is then necessary that the parts of the distributed transmitter could code a message using rotated and scaled Alamouti. The virtual (or distributed) transmitter occurs e.g. in relay communication. The real transmitter sends a message to two relays, these two relays then act as one distributed transmitter. In other words, the encoding and transmission could be done at different locations. So instead of one transmitter with two antennas, there could be two (cooperating) transmitters each having one antenna.

[0104] Different from the above examples, the scaling function and/or the rotation function could be applied to different symbols than the symbols for which they are applied in these examples. The scaling function and/or the rotation function could be, for instance, be applied to all symbols. According to the invention the mapping should be such that it is piecewise linear. The scaling function scales the incoming signal with a constant that is typically 2, 3, 4, etc. but also complex scaling factor are possible e.g. 2+j.

[0105] Further, also a single receive antenna could be used according to the present invention.

[0106] As explained above the decoding method is different from the Alamouti decoder/receiver. According to the Alamouti method the received signals are orthogonal so the symbol estimates with no noise enhancement can be separated and the optimum decoding (ML) is very simple to implement. In the scaled-repeated code this is not the case and the optimum decoding is more complex. However, given the properties of the chosen scaling function, according to the present invention a simple sub-optimum decoding method was derived which basically performs as good as the optimum decoding.

[0107] In a general sense, the decoding has to match the encoding. So, if the code is changed, also the decoding will change accordingly. However, as far as the condition on the scaling M(M(x))=-x is satisfied, the decoding has similar possible sub-optimum and simple structure.

[0108] Briefly summarized the decoding according to the present invention consists generally of the following steps:

1) Check (46) and decide to decode (s_1,s_2) or its scaled version (t_1,t_2).

2) For each alternative of $((D_2(s_1),D_2(s_2))$ (This checking for couples comes from the piece-wise linear with at least two pieces. There are 16 possible couples for $((D_2(s_1),D_2(s_2))$ since $D_2()$ returns the complex sign of its argument which can have the following four values (+1+j, +1-j, -1-j, -1+j). For $D_3()$, instead of the sign function 3 regions are defined for the real dimension and 3 regions for the imaginary dimension (see (56)). Therefore 3*3x3*3=81 possible couples are obtained.):

   2a) Use equations (49)-(52) to derive an estimate $(\tilde{s}_1,\tilde{s}_2)$ of the couple (s_1,s_2). (The vectors *c,d* and *c',d'* remain the same independently of the constellation size. The vector *a,b* and *a',b'* changes a bit. Using M_2, they have the factor 2 in some of the elements. Using M_3, they have the factor 3 instead of the factor 2.)

   2b) Calculate and save the Euclidean distance between the received vector $\underline{z}$ and the estimate $(\tilde{s}_1\underline{a}+\tilde{s}_2\underline{b})$ (see (53)

   2c) Slice the estimate of (s_1,s_2) and save the sliced results

3) The final estimate of (s_1,s_2) is the sliced result corresponding to the minimum Euclidean distance.

[0109] In Method 1 and Method 2, the step 2c) is performed only if the current Euclidean distance is the smallest one encountered so far.

[0110] Method 2 further reduces the number of slicing steps, by starting the investigation of pairs $((D_2(s_1),D_2(s_2))$ from *y* a good guess of $((D_2(s_1),D_2(s_2))$ obtained by projecting the received vector z of equation (49) onto the space defined by e_1 and e_2 of (55). (Also e_1 and e_2 change depending on the scaling function. With M_2, they are defined in (55). With M_3, the 1/2 factor in e_1 and e_2 changes in a 1/3 factor.)

[0111] The slicer depends on the constellation size as it does in a regular communication system. A 4QAM and a 16-QAM have different slicer.

[0112] In a another embodiment of the invention, it has been noticed that current systems use M-QAM constellations that take integer values and are not designed for diversity transmission.

[0113] They are designed to increase the minimum Euclidian distance (main factor that determined the SER), but the minimum product distance is not considered. According to this other embodiment, new constellations are proposed. These constellations do not need to take integer values, and should consider the minimum product distance as the design criteria.

[0114] Consider also the min Euclidian distance for the worst-case scenarios where the channels are the same. The Euclidian distance criteria imples a Repetition structure (Co-Re) that should look like QAM-like constellation in diversity branch dimensions. Any rotated version of conventional M-QAM constellations in 2 branch dimensions achieves the same Euclidian distance.

[0115] The optimum rotation in branch dimensions should maximize the minimum product distance. The requirement is n bits/ real dimension. A conventional constellation is 2^(2n) -QAM /complex dimension.

[0116] According to this new constellation it is porposed the following:

- Step 1: Start with 2^n -QAM constellation in diversity branch dimensions (x=transmitted value in branch 1, y=transmitted value in branch 2, z =x+j*y $\in$ 2^n QAM);
- Step 2: Rotate 2^n QAM with $\theta=1/2*\tan^{-1}(2)$ and z $\in$ 2^n QAM, thus x=Real(z*exp(j θ)), y=Imag(z*exp(j θ)), x or y set will form the new PAM constellation / real dimension (T) and x and y will form the mapping structure of Co-Re scheme for each real dimension.
- Step 3: Form the complex constellation by using two PAM obtained by T for each dimension:

s1 $\in$ New constellation: Cnew={s1=x1+j*x2 | x1, x2 $\in$ T} for the first transmission.
s2 $\in$ New constellation: Cnew={s2=y1+j*y2 | y1, y2 $\in$ T} for the first transmission.

[0117] In an example of this embodiment, the optimum rotation angle that maximizes the minimum product distance is θ=1/2*tan-1(2). Thus, these new constellation design for 2 branch diversity systems enable an improved minimum product distance, leading to a lower SER with Co-Re. Morover, this improvement is higher in higher constellation sizes. This is applicable to any 2-branch diversity scheme, e.g., WLAN, cellular, broadcast or sensor networks, and for instance for channel estimation in (wireless) systems. The tx signal includes some known pilot sequences so that the rx can estimate how the signal has been corrupted by the channel. DVB-T also performs a channel estimation but in presence of a SFN and using OFDM modulation. SFN makes the channel very long. OFDM allows for the insertion of pilots in the time-frequency grid.

[0118] This embodiment has several advantages as: a good minimum product distance, a good symbol-error-rate performance, is easily scalable to higher constellations sizes, starts with conventional QAM constellations, and rotates with the optimum angle. However, the new constellations are not on the grid, and it is hard to get the exact values. Thus, it Requires more memory for each constellation point and makes the slicing operation more complex. In a variant of this embodiment, different constellations are proposed. The constellations points are on the grid, i.e., take integer values as conventional QAM, and may be non-uniformly distributed, i.e., not all points in the constellation are equidistant to each other. According to a method pursuant to this variant, first step is considering all possible integer values for each constellation size. Then, normalizing the average power and comparing the constellations in terms of minimum product distance. For instance, a 4-PAM constellation design (Ordinary 4-PAM dproduct=64, optimum 4-PAM dproduct=80.17) Options:

a) [-2 -1 1 2] -> [-2 -1 1 2] *squareroot(10)/squareroot(5) ->dproduct=36.
b) [-4 -1 1 4]-> [-4 -1 1 4] *squareroot(10)/squareroot(17)-> dproduct=77.85
c) [-5 -1 1 5]-> [-5 -1 1 5] *squareroot(10)/squareroot(26) -> dproduct=59.17
d) [-5 -2 2 5]-> [-5 -2 2 5] *squareroot(10)/squareroot(29) -> dproduct=52.44... [-4 -1 1 4] 4-PAM constellation seems to be a good choice that has a high minimum product distance and it is on the grid. Complex constellations can be devised by using the new PAM constellations in both real and imaginary axes. 4-PAM -> new 16 QAM complex constellation. For larger constellation sizes:

Option 1: Same approach can be applied, i.e., search every possibility considering the minimum product distance as the criteria
Option 2: Use the basic structure of one of the new 4-PAM constellations and repeat it. Option 2 seems to be more practical

For instance, designing new 16-PAM (256 QAM) constellations by replicating the [-4 -1 1 4] structure to obtain a new 16-PAM. These constellations on the grid for 2 branch diversity systems enables good minimum product distance-> Low MER with Co-Re. The improvement is higher in higher constellation sizes, still on the grid, i.e., take integer values which requires less memory in the receiver. This imporvement is applicable to any 2-branch diversity scheme, e.g., WLAN, cellular, broadcast or sensor networks, STBCs, Space-time Trellis Codes, OTD.

[0119] In another embodiment of the invention, it is proposed a new Space time block coding (STBC) for 2 transmit antenna systems using the concepts of scaled repetition and rotation but without any conjugation of the symbols as in Alamouti coding. Especially in high constellation sizes, the new STBC structure provides better symbol-error-rate performance than STBCs and achieves the same minimum determinant as the Golden Codes and a similar SER performance. RSA is based on scaled repetition in Alamouti coding. Alamouti coding provides a simplified receiver structure for ordinary repetition. Better performance-complexity trade-offs can be achieved if we do not restrict to Alamouti structure. Min determinant criteria (c and c' are any possible pair of STBC codes)

$$\Pr\{\underline{c} \to \underline{c}'\} \approx \gamma' \frac{1}{\det^2((\underline{c}' - \underline{c})(\underline{c}' - \underline{c})^{\dagger})} \frac{1}{\mathsf{SNR}^4}.$$

Min determinant is highly dependent on the minimum product distance of the constellation-rearrangement scheme, thus, use scaled repetition scheme with the constellations defined in the previous embodiment. New scheme may help to increase the minimum determinant by modifying the existing STBC structure.

[0120] Since the product distance of the streams are affected by the new STBC structure, the optimum rotation angle that maximizes the minimum determinant has changed.

Optimum rotation is $\theta=\pi/2$ for all constellation sizes. Same minimum determinant as the Golden code for all constellation sizes. Min absolute value of determinant is 8.9443 for 16 QAM. The min determinant of Golden Code with this rate is also 8.9443.

[0121] For R=6.34 bits/transm. constellation size
Increased min determinant-> Similar MER with the Golden Code
Much better performance than the other competitors, e.g., Alamouti, uncoded, tilted QAM or RSA. A new STBC structure for 2 transmit antenna systems
This structure permits a higher min determinant, a more efficient space-time block codes, a better SER performance than RSA coding, with a better utilization of scaled repetition approach in designing STBCs. A similar SER performance with the Golden Codes.
In accordance with another variant of the invetion, a new space-time block coding structure, Rotated and Scaled Alamouti coding (RSA), is proposed for two transmit antenna systems. It is shown that the new STBC outperforms the well-known Alamouti coding, and provides a robust transmission scheme without requiring a very complex receiver structure. It is also mentioned that there exist some other competitor space-time block codes, i.e., Golden codes, that also outperform the Alamouti coding. The proposed RSA coding performs slightly worse than the Golden code. However, it enjoys a simpler decoding mechanism than the Golden code which uses exhaustive ML search for decoding. This variant aims to improve the performance of the RSA coding by providing a new repetition structure utilizing the new constellations and constellation re-arrangement scheme designed for 2-branch diversity systems proposed previously. By applying an appropriate rotation, the new repetition structure provides the RSA coding a better symbol-error-rate performance than the current repetition structure and reduces the performance gap between RSA coding and Golden code. For certain constellation sizes it completely eliminates the performance gap. The RSA coding still enjoys the simple receiver structure according to the invention.
Scaled repetition scheme is based on conventional Co-Re schemes. In the previous variants of the invention, we showed that the performance of conventional Co-Re schemes can be improved by using new constellations and new Co-Re structures. These new constellations and Co-Re structures can be used to increase the min determinant criteria that determines the symbol error rate of the STBC: Min determinant criteria (c and c' are any possible pair of STBC codes).

$$\Pr\{\underline{c} \to \underline{c}'\} \approx \gamma' \frac{1}{\det^2((\underline{c}' - \underline{c})(\underline{c}' - \underline{c})^{\dagger})} \frac{1}{\mathrm{SNR}^4}.$$

Min determinant is highly dependent on the minimum product distance of the constellation-rearrangement scheme (scaled repetition scheme in RSA)
For RSA,

$$\det(c'-c) = e^{j\theta}\underbrace{(s_1'-s_1)((M_2(s_1'^*)-M_2(s_1^*))}_{\text{product distance}} + \underbrace{(s_2'^*-s_2^*)((M_2(s_2')-M_2(s_2)))}_{\text{product distance}}$$

[0122] New constellations and constellation re-arrangement scheme may help to increase the minimum determinant by modifying the existing scaled repetition scheme. New scaled repetition scheme for RSA coding using new constellations and constellation re-arrangement.
It changes only the repetition structure, not the scaling concept. The optimum rotation angle that maximizes the minimum product distance is found as δ =1/2*tan^{-1}(2). Scaling factor=tan(δ)=(1-root5)/2. This scaling factor is fixed for all constellation sizes.
Since the product distance of the streams are affected by the new scaled repetition scheme, the optimum rotation angle that maximizes the minimum determinant has changed. Optimum rotation is θ=tan^{-1}(2) for 16-QAM(4-PAM for each real dimension).
Min absolute value of determinant is 8.9443> 7.613
The min determinant of Golden Code with this rate is also 8.9443.
For R=4 bits/transmitter constellation size. An increased min determinant leads to similar SER with the Golden Code. It still enjoys the simple receiver mechanism and much better performance than the other competitors e.g., Alamouti, uncoded, tilted QAM.
For higher constellation sizes, the performance gap between Golden Code and RSA will be reduced by using the new repetition structure while RSA will still enjoy the simpler decoding structure.

[0123] Recently, Golden Code is proposed for two transmit antenna systems. It is shown that this new space-time block code outperforms the well-known Alamouti coding, and provides a robust transmission scheme. However, it requires a very complex receiver structure, i.e., exhaustive ML search. This variant of the invention proposes a suboptimum low complexity receiver structure for decoding the Golden Code by using the same approach used in detecting the RSA coding according to the invention. Especially in high constellation sizes, the new receiver structure provides a substantial reduction in the computational complexity while providing an acceptable level of symbol-error-rate close to the ML detection.

Golden Code Structure [1]

[0124]

$$\begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} = \frac{1}{\sqrt{5}} \begin{pmatrix} \alpha(\mathrm{a}+\theta\,\mathrm{b}) & \alpha(\mathrm{c}+\theta\,\mathrm{d}) \\ j\bar{\alpha}(\mathrm{c}+\bar{\theta}\,\mathrm{d}) & \bar{\alpha}(\mathrm{a}+\bar{\theta}\,\mathrm{b}) \end{pmatrix}$$

[0125] Where,

- *a*, *b*, *c* and *dare* M - QAM symbols
- $\bar{\theta} = 1 - \theta = \frac{1-\sqrt{5}}{2}$, $\alpha = 1 + j(1-\theta)$ and $\overline{\alpha} = 1+7(1-\overline{\theta})$

[0126] Golden Code can be viewed as space-time coding scheme based on scaled repetition Scaled repetition provides two different interpretations of the transmitted signal Different interpretation means different spatial signatures Choose the best interpretation (set of spatial signatures) to apply zero-forcing (ZF) receivers: the lowest noise enhancement with ZF receivers and check the Euclidian distance for each sub-region.

Interpretation of the Golden Code:

[0127]

$$\frac{1}{\sqrt{5}} \begin{pmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{pmatrix} = \frac{1}{\sqrt{5}} \begin{pmatrix} \alpha(a+\theta\, b) & \alpha(c+\theta\, d) \\ j\bar{\alpha}(c+\bar{\theta}\, d) & \bar{\alpha}(a+\bar{\theta}\, b) \end{pmatrix}$$

[0128] For a given $a = a_j$,

$$x_{22} = \bar{\alpha} a_j + \bar{\alpha}\bar{\theta}\,[\frac{x_{11} - \alpha a_j}{\alpha\theta}] = \underbrace{\frac{\bar{\alpha}\bar{\theta}}{\alpha\theta}}_{\beta_1:\text{Scaling Factor}} x_{11} + \underbrace{\bar{\alpha}(1 - \frac{\bar{\theta}}{\theta}) a_j}_{D_1(a_j):\text{Offset}}$$

$x_{11} = \alpha\theta b + \alpha a_j$, $x_{22} = \overline{\alpha} a_j + \overline{\alpha\theta} b \Rightarrow$

[0129] Similarly, for a given b b_i,

$$x_{11} = \alpha\theta\, b_i + \alpha\,[\frac{x_{22} - \bar{\alpha}\bar{\theta}\, b_i}{\bar{\alpha}}] = \underbrace{\frac{\alpha}{\bar{\alpha}}}_{\beta_2:\text{Scaling Factor}} x_{22} + \underbrace{\alpha(\theta - \bar{\theta}) b_i}_{D_2(b_i):\text{Offset}}$$

$x_{11} = \alpha\theta\, b_i + \alpha a$, $x_{22} = \overline{\alpha} a + \overline{\alpha\theta}\, b_i \Rightarrow$

[0130] There are four different Interpretation of the Golden Code:

$$\begin{pmatrix} x_{11} & x_{21} \\ x_{12} & x_{22} \end{pmatrix} = \begin{pmatrix} \alpha(a+\theta\, b) & \alpha(c+\theta\, d) \\ j\overline{\alpha}(c+\overline{\theta}\, d) & \overline{\alpha}(a+\overline{\theta}\, b) \end{pmatrix}$$

$$= \begin{pmatrix} x_{11} & x_{21} \\ j\beta_1 x_{21} & \beta_1 x_{11} \end{pmatrix} + \begin{pmatrix} 0 & 0 \\ jD_1(c) & D_1(a) \end{pmatrix} \qquad (1)$$

$$= \begin{pmatrix} \beta_2 x_{22} & x_{21} \\ j\beta_1 x_{21} & x_{22} \end{pmatrix} + \begin{pmatrix} D_2(b) & 0 \\ jD_1(c) & 0 \end{pmatrix} \qquad (2)$$

$$= \begin{pmatrix} x_{11} & -j\beta_2 x_{12} \\ x_{12} & \beta_1 x_{11} \end{pmatrix} + \begin{pmatrix} 0 & D_2(d) \\ 0 & D_1(a) \end{pmatrix} \qquad (3)$$

$$= \begin{pmatrix} \beta_2 x_{22} & -j\beta_2 x_{12} \\ x_{12} & x_{22} \end{pmatrix} + \begin{pmatrix} D_2(b) & D_2(d) \\ 0 & 0 \end{pmatrix} \qquad (4)$$

Leading to 4 Different Spatial Signature Sets

[0131] The received signal in a 2x2 MIMO system is

$$\mathbf{r_1} = \frac{1}{\sqrt{5}} \begin{pmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \end{pmatrix} \begin{pmatrix} x_{11} \\ x_{21} \end{pmatrix} + \begin{pmatrix} n_{11} \\ n_{12} \end{pmatrix},$$

in the first time slot

$$\mathbf{r}_2 = \frac{1}{\sqrt{5}} \begin{pmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \end{pmatrix} \begin{pmatrix} x_{12} \\ x_{22} \end{pmatrix} + \begin{pmatrix} n_{21} \\ n_{22} \end{pmatrix},$$

in the second time slot

[0132] Thus,

$$\begin{pmatrix} \mathbf{r_1} \\ \mathbf{r}_2 \end{pmatrix} = \frac{1}{\sqrt{5}} \begin{pmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \\ \beta_1 h_{21} & j\beta_1 h_{11} \\ \beta_1 h_{22} & j\beta_1 h_{12} \end{pmatrix} \begin{pmatrix} x_{11} \\ x_{12} \end{pmatrix} + \mathbf{M}(a,c) + \begin{pmatrix} \mathbf{n_1} \\ \mathbf{n_2} \end{pmatrix} = \frac{1}{\sqrt{5}} \begin{pmatrix} \beta_2 h_{11} & h_{21} \\ \beta_2 h_{12} & h_{22} \\ h_{21} & j\beta_1 h_{11} \\ h_{22} & j\beta_1 h_{12} \end{pmatrix} \begin{pmatrix} x_{22} \\ x_{12} \end{pmatrix} + \mathbf{M}(b,c) + \begin{pmatrix} \mathbf{n_1} \\ \mathbf{n_2} \end{pmatrix}$$

$$= \frac{1}{\sqrt{5}} \begin{pmatrix} h_{11} & -j\beta_2 h_{21} \\ h_{12} & -j\beta_2 h_{22} \\ \beta_1 h_{21} & h_{11} \\ \beta_1 h_{22} & h_{12} \end{pmatrix} \begin{pmatrix} x_{11} \\ x_{21} \end{pmatrix} + \mathbf{M}(a,d) + \begin{pmatrix} \mathbf{n_1} \\ \mathbf{n_2} \end{pmatrix} = \frac{1}{\sqrt{5}} \begin{pmatrix} \beta_2 h_{11} & -j\beta_2 h_{21} \\ \beta_2 h_{12} & -j\beta_2 h_{22} \\ h_{21} & h_{11} \\ h_{22} & h_{12} \end{pmatrix} \begin{pmatrix} x_{21} \\ x_{22} \end{pmatrix} + \mathbf{M}(b,d) + \begin{pmatrix} \mathbf{n_1} \\ \mathbf{n_2} \end{pmatrix}$$

[0133] It is thus proposed a suboptimum Receiver Structure:

Golden Code can be interpreted in 4 different ways
4 Different Spatial Signature Sets
The simplest form of receiver is zero-forcing (ZF)
Apply the ZF receiver to the best interpretation
Search for lowest noise enhancement

% of the signal power lost due to ZF receiver

$$P_{loss,i} = \frac{|s_{1(i)}{}^{H} s_{2(i)}|}{|s_{1(i)}|\,|s_{2(i)}|} \quad \text{where}\, \mathbf{S_i} = [s_{1(i)} \;\; s_{2(i)}]$$

[0134] According to a method pursuant to this example, a first step is finding the Si that minimizes the loss, then applying the ZF receiver of Si and checking all possible offset values, i.e., M.

[0135] The following table is given here for comparison of the complexity of these systems.

Modulation	Optimum Detection	Proposed Detection	ZF Receiver
BPSK	16 exhaustive search (Euclidian distance computation)	8 Slicing	4 Slicing
QPSK	256 exhaustive search (Euclidian distance computation)	32 Slicing	4 Slicing
16QAM	65536 exhaustive search (Euclidian distance computation)	512 Slicing	4 Slicing

[0136] A low complexity Golden Code Decoder is proposed based on different interpretations of the Golden Code. The receiver is simpler than ML detection but more complex than ZF receiver. Thus, is a good trade-off between complexity and SER performance.

[0137] While the invention has been illustrated and described in detail in the drawings and foregoing description, such illustration and description are to be considered illustrative or exemplary and not restrictive; the invention is not limited to the disclosed embodiments.

[0138] Other variations to the disclosed embodiments can be understood and effected by those skilled in the art in practicing the claimed invention, from a study of the drawings, the disclosure, and the appended claims.

[0139] In the claims, the word "comprising" does not exclude other elements or steps, and the indefinite article "a" or "an" does not exclude a plurality. A single unit may fulfill the functions of several items recited in the claims. The mere fact that certain measures are recited in mutually different dependent claims does not indicate that a combination of these measured cannot be used to advantage.

[0140] A computer program may be stored/distributed on a suitable medium, such as an optical storage medium or a solid-state medium supplied together with or as part of other hardware, but may also be distributed in other forms, such as via the Internet or other wired or wireless telecommunication systems.

[0141] Any reference signs in the claims should not be construed as limiting the scope.

Claims

1. Encoder for encoding incoming symbols of an incoming data stream into channel symbols of a channel data stream for transmission over a transmission channel **characterized by** comprising:

   - mapping means for block by block mapping incoming symbols onto pairs of channel symbols, a block comprising two incoming symbols, the mapping means being arranged for mapping the block onto two pairs of channel symbols such that said two pairs of channel symbols include scaled versions of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols, said scaled versions being obtained by applying a scaling function having a scaling factor with an absolute value different from one and being piece-wise linear with at least two pieces, and
   - output means for outputting said channel symbols.

2. Encoder as claimed in claim 1,
   wherein said mapping means are adapted for applying a rotation function for rotating at least one of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols by a rotation angle such that said two pairs of channel symbols include a rotated version of at least one of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols, wherein said rotation angle is different from 0 and 180 degrees.

3. Encoder as claimed in claim 2,

wherein said mapping means are adapted for applying a rotation function for rotating at least one of said two incoming symbols by a predetermined rotation angle, that is chosen to maximize the minimum modulus of the determinant of the code matrices.

4. Encoder as claimed in claim 2,
wherein said mapping means are adapted for applying a rotation function for rotating at least one of said two incoming symbols by a fixed predetermined rotation angle.

5. Encoder as claimed in claim 2,
wherein said mapping means are adapted for applying a rotation function for rotating a first of said two incoming symbols by a rotation angle such that a first pair of channel symbols includes said rotated version of said first incoming symbol.

6. Encoder as claimed in claim 2,
wherein said mapping means are adapted for block by block mapping two incoming symbols onto two pairs of channel symbols such that
a first pair of channel symbols includes a rotated version of one of the two incoming symbols and a scaled version of the other incoming symbol and
a second pair of channel symbols includes a scaled version of said one of the two incoming symbols and a negated and complex conjugate version of said other incoming symbol.

7. Transmitter (10) for transmitting channel symbols of a channel data stream over a transmission channel comprising:

- an encoder as claimed in claim 1 for encoding incoming symbols of an incoming data stream into channel symbols of said channel data stream, and
- transmission means, in particular two transmission antennas (11, 12), for receiving said channel symbols from said encoder and for transmitting said channel symbols over said transmission channel.

8. Encoding method for encoding incoming symbols of an incoming data stream into channel symbols of a channel data stream for transmission over a transmission channel, the encoding method being **characterized by** comprising the steps of:

- block by block mapping incoming symbols onto pairs of channel symbols, a block comprising two incoming symbols and being mapped onto two pairs of channel symbols such that said two pairs of channel symbols include scaled versions of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols, said scaled versions being obtained by applying a scaling function having a scaling factor with an absolute value different from one and being piece-wise linear with at least two pieces, and
- outputting said channel symbols.

9. Decoder being adapted for block by block decoding received channel symbols of a channel data stream, which have been encoded from incoming symbols of an incoming data stream, and transmitted over a transmission channel, wherein during encoding incoming symbols have been block by block mapped onto pairs of channel symbols, a block comprising two incoming symbols and being mapped onto two pairs of channel symbols such that said two pairs of channel symbols include scaled versions of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols, said scaled versions being obtained by applying a scaling function having a scaling factor with an absolute value different from one and being piece-wise linear with at least two pieces, said decoder further comprising:

- selection means for selecting a pair of possible function values of incoming symbols s for decoding a current block of received channel symbols comprising a pair of received channel symbols y and,
- subtraction means for determining first intermediate scaled versions $D_2(s)$ of said selected pair of possible function values of incoming symbols s by applying a sub-function D_2 and for subtracting terms including said first intermediate scaled versions $D_2(s)$ of said selected pair of possible function values of incoming symbols s from said pair of received channel symbols y to obtain second intermediate versions z of said selected pair of possible function values of incoming symbols s,
- detection means for detecting third intermediate versions $\tilde{s}$ of said selected pair of possible function values of incoming symbols s by applying a zero force detection,
- calculating means for calculating the Euclidian distance between the received signal and the estimated symbols,

- slicing means for slicing said estimate, third intermediate versions $\tilde{s}$ of said selected pair of possible function values of incoming symbols s to obtain estimates of said selected pair of possible function values of incoming symbols s, and
- control means for repeating said steps with other pairs of possible incoming symbols s until a predetermined stop condition is met or until a minimum Euclidian distance is found and for outputting said pair of possible incoming symbols s resulting in the minimum Euclidian distance.

10. Receiver (20, 50) for receiving channel symbols of a channel data stream over a transmission channel comprising:

- receiving means, in particular one or two receiving antennas (21, 51, 52), for receiving said channel data stream over said transmission channel and for outputting the channel symbols said channel data signal to a decoder, and
- a decoder as claimed in claim 9 for block by block decoding the received channel symbols of said channel data stream.

11. Encoded data signal for transmission over a transmission channel carrying channel symbols of a channel data stream, which have been encoded from incoming symbols of an incoming data stream, wherein during encoding incoming symbols have been block by block mapped onto pairs of channel symbols, a block comprising two incoming symbols and being mapped onto two pairs of channel symbols such that said two pairs of channel symbols include scaled versions of said two incoming symbols and/or of the complex conjugate of at least one of said two incoming symbols, said scaled versions being obtained by applying a scaling function having a scaling factor with an absolute value different from one and being piece-wise linear with at least two pieces.

12. Computer program product comprising program code means for causing a computer to carry out the steps of the method as claimed in claim 8, when said computer program is carried out on a computer.

Patentansprüche

1. Codierer zur Codierung eingehender Symbole eines eingehenden Datenstroms in Kanalsymbole eines Kanaldatenstroms zur Übertragung über einen Übertragungskanal, **dadurch gekennzeichnet, dass** er umfasst:

- Zuordnungsmittel, um eingehende Symbole Paaren von Kanalsymbolen blockweise zuzuordnen, wobei ein Block zwei eingehende Symbole umfasst, wobei die Zuordnungsmittel so eingerichtet sind, dass sie den Block zwei Paaren von Kanalsymbolen so zuordnen, dass die beiden Kanalsymbolpaare skalierte Versionen der zwei eingehenden Symbole und/oder der komplex Konjugierten von mindestens einem der zwei eingehenden Symbole enthalten, wobei die skalierten Versionen durch Anwenden einer Skalierungsfunktion erhalten werden, die einen Skalierungsfaktor mit einem absoluten Wert anders als Eins aufweist und mit mindestens zwei Stück stückweise linear ist, sowie
- Ausgabemittel zur Ausgabe der Kanalsymbole.

2. Codierer nach Anspruch 1,
wobei die Zuordnungsmittel so eingerichtet sind, dass sie eine Rotationsfunktion anwenden, um mindestens eines der zwei eingehenden Symbole und/oder der komplex Konjugierten von mindestens einem der zwei eingehenden Symbole um einen Rotationswinkel so zu drehen, dass die beiden Paare von Kanalsymbolen eine gedrehte Version von zumindest einem der beiden eingehenden Symbole und/oder der komplex Konjugierten von zumindest einem der zwei eingehenden Symbole enthalten, wobei der Rotationswinkel von 0 und 180 Grad verschieden ist.

3. Codierer nach Anspruch 2,
wobei die Zuordnungsmittel so eingerichtet sind, dass sie eine Rotationsfunktion anwenden, um mindestens eines der zwei eingehenden Symbole um einen vorgegebenen Rotationswinkel zu drehen, der so gewählt wird, dass der Mindestmodul der Determinante der Codematrizes maximiert wird.

4. Codierer nach Anspruch 2,
wobei die Zuordnungsmittel so eingerichtet sind, dass sie eine Rotationsfunktion anwenden, um mindestens eines der zwei eingehenden Symbole um einen festen vorgegebenen Rotationswinkel zu drehen.

5. Codierer nach Anspruch 2,

wobei die Zuordnungsmittel so eingerichtet sind, dass sie eine Rotationsfunktion anwenden, um ein erstes der zwei eingehenden Symbole so um einen Rotationswinkel zu drehen, dass ein erstes Paar von Kanalsymbolen die gedrehte Version des ersten eingehenden Symbols enthält.

6. Codierer nach Anspruch 2,
wobei die Zuordnungsmittel so eingerichtet sind, dass sie zwei eingehende Symbole zwei Paaren von Kanalsymbolen blockweise so zuordnen, dass
ein erstes Paar von Kanalsymbolen eine gedrehte Version von einem der zwei eingehenden Symbole sowie eine skalierte Version des anderen eingehenden Symbols enthält, und
ein zweites Paar von Kanalsymbolen eine skalierte Version des einen der zwei eingehenden Symbole sowie eine negierte und konjugiert komplexe Version des anderen eingehenden Symbols enthält.

7. Sender (10) zur Übertragung von Kanalsymbolen eines Kanaldatenstroms über einen Übertragungskanal, umfassend:

- einen Codierer nach Anspruch 1 zur Codierung eingehender Symbole eines eingehenden Datenstroms in Kanalsymbole des Kanaldatenstroms, sowie
- Übertragungsmittel, insbesondere zwei Sendeantennen (11, 12), um von dem Codierer die Kanalsymbole zu empfangen und die Kanalsymbole über den Übertragungskanal zu übertragen.

8. Codierungsverfahren zur Codierung eingehender Symbole eines eingehenden Datenstroms in Kanalsymbole eines Kanaldatenstroms zur Übertragung über einen Übertragungskanal, wobei das Codierungsverfahren durch die folgenden Schritte gekennzeichnet ist, wonach:

- eingehende Symbole Paaren von Kanalsymbolen blockweise zugeordnet werden, wobei ein Block zwei eingehende Symbole umfasst und zwei Paaren von Kanalsymbolen so zugeordnet wird, dass die beiden Kanalsymbolpaare skalierte Versionen der zwei eingehenden Symbole und/oder der komplex Konjugierten von mindestens einem der zwei eingehenden Symbole enthalten, wobei die skalierten Versionen durch Anwenden einer Skalierungsfunktion erhalten werden, die einen Skalierungsfaktor mit einem absoluten Wert anders als Eins aufweist und mit mindestens zwei Stück stückweise linear ist, und
- die Kanalsymbole ausgegeben werden.

9. Decodierer, der so eingerichtet ist, dass er empfangene Kanalsymbole eines Kanaldatenstroms decodiert, die aus eingehenden Symbolen eines eingehenden Datenstroms codiert und über einen Übertragungskanal übertragen wurden, wobei während des Codierens eingehende Symbole Paaren von Kanalsymbolen blockweise zugeordnet wurden, wobei ein Block zwei eingehende Symbole umfasst und zwei Paaren von Kanalsymbolen so zugeordnet wird, dass die beiden Kanalsymbolpaare skalierte Versionen der zwei eingehenden Symbole und/oder der komplex Konjugierten von mindestens einem der zwei eingehenden Symbole enthalten, wobei die skalierten Versionen durch Anwenden einer Skalierungsfunktion erhalten werden, die einen Skalierungsfaktor mit einem absoluten Wert anders als Eins aufweist und mit mindestens zwei Stück stückweise linear ist, wobei der Decodierer weiterhin umfasst:

- Auswahlmittel, um ein Paar möglicher Funktionswerte von eingehenden Symbolen s zur Decodierung eines aktuellen Blocks empfangener Kanalsymbole mit einem Paar von empfangenen Kanalsymbolen y auszuwählen, sowie
- Subtraktionsmittel, um erste skalierte Zwischenversionen $D_2(s)$ des ausgewählten Paares möglicher Funktionswerte von eingehenden Symbolen s durch Anwenden einer Subfunktion D_2 zu ermitteln und Terme mit den ersten skalierten Zwischenversionen $D_2(s)$ des ausgewählten Paares möglicher Funktionswerte von eingehenden Symbolen s von dem Paar von empfangenen Kanalsymbolen y zu subtrahieren, um zweite Zwischenversionen z des ausgewählten Paares möglicher Funktionswerte von eingehenden Symbolen s zu erhalten,
- Detektionsmittel, um dritte Zwischenversionen $\tilde{s}$ des ausgewählten Paares möglicher Funktionswerte von eingehenden Symbolen s durch Anwenden einer Nullkraftdetektion zu detektieren,
- Berechnungsmittel zum Berechnen der Euklidischen Distanz zwischen dem empfangenen Signal und den geschätzten Symbolen,
- Slicing-Mittel zum Slicen der dritten Schätz-Zwischenversionen $\tilde{s}$ des ausgewählten Paares möglicher Funktionswerte von eingehenden Symbolen s, um Schätzungen des ausgewählten Paares möglicher Funktionswerte von eingehenden Symbolen s zu erhalten, sowie
- Steuermittel, um die Schritte mit anderen Paaren von möglichen eingehenden Symbolen s zu wiederholen, bis eine vorgegebene Stopp-Bedingung erfüllt wird oder bis eine minimale Euklidische Distanz ermittelt wird,

und um das Paar von möglichen eingehenden, in der minimalen Euklidischen Distanz resultierenden Symbolen s auszugeben.

10. Empfänger (20, 50) zum Empfang von Kanalsymbolen eines Kanaldatenstroms über einen Übertragungskanal, umfassend:

- Empfangsmittel, insbesondere eine oder zwei Empfangsantennen (21, 51, 52), um den Kanaldatenstrom über den Übertragungskanal zu empfangen und um die Kanalsymbole des Kanaldatenstroms an einen Decodierer auszugeben; sowie
- einen Decodierer nach Anspruch 9, um die empfangenen Kanalsymbole des Kanaldatenstroms blockweise zu decodieren.

11. Codiertes Datensignal zur Übertragung über einen Übertragungskanal, der Kanalsymbole eines Kanaldatenstroms leitet, die aus eingehenden Symbolen eines eingehenden Datenstroms codiert wurden, wobei während des Codierens eingehende Symbole Paaren von Kanalsymbolen blockweise zugeordnet wurden, wobei ein Block zwei eingehende Symbole umfasst und zwei Paaren von Kanalsymbolen so zugeordnet wird, dass die beiden Kanalsymbolpaare skalierte Versionen der zwei eingehenden Symbole und/oder der komplex Konjugierten von mindestens einem der zwei eingehenden Symbole enthalten, wobei die skalierten Versionen durch Anwenden einer Skalierungsfunktion erhalten werden, die einen Skalierungsfaktor mit einem absoluten Wert anders als Eins aufweist und mit mindestens zwei Stück stückweise linear ist.

12. Computerprogrammprodukt mit Programmcodemitteln, um einen Computer dazu zu veranlassen, die Schritte des Verfahrens nach Anspruch 8 auszuführen, wenn das Computerprogramm auf einem Computer ausgeführt wird.

Revendications

1. Codeur pour coder des symboles entrants d'un courant de données entrant en symboles de canal d'un courant de données de canal pour transmission via un canal de transmission, **caractérisé en ce qu'**il comprend :

- des moyens de mappage pour mapper bloc par bloc des symboles entrants en paires de symboles de canal, un bloc comprenant deux symboles entrants, les moyens de mappage étant agencés pour mapper le bloc en deux paires de symboles de canal de sorte que lesdites deux paires de symboles de canal comprennent des versions cadrées desdits deux symboles entrants et/ou du conjugué complexe d'au moins l'un desdits deux symboles entrants, lesdites versions cadrées étant obtenues en appliquant une fonction de cadrage ayant un facteur de cadrage avec une valeur absolue différente de un et qui est linéaire pièce par pièce avec au moins deux pièces, et
- des moyens de sortie pour délivrer lesdits symboles de canal.

2. Codeur selon la revendication 1,
dans lequel lesdits moyens de mappage sont adaptés pour appliquer une fonction de rotation pour faire tourner au moins l'un desdits deux symboles entrants et/ou du conjugué complexe d'au moins l'un desdits deux symboles entrants d'un angle de rotation tel que lesdites deux paires de symboles de canal comprennent une version tournée d'au moins l'un desdits deux symboles entrants et/ou du conjugué complexe d'au moins l'un desdits deux symboles entrants, dans lequel ledit angle de rotation est différent de 0 et 180 degrés.

3. Codeur selon la revendication 2,
dans lequel lesdits moyens de mappage sont adaptés pour appliquer une fonction de rotation pour faire tourner au moins l'un desdits deux symboles entrants d'un angle de rotation prédéterminé qui est choisi pour maximiser le module minimal du déterminant des matrices de code.

4. Codeur selon la revendication 2,
dans lequel lesdits moyens de mappage sont adaptés pour appliquer une fonction de rotation pour faire tourner au moins l'un desdits deux symboles entrants d'un angle de rotation prédéterminé fixe.

5. Codeur selon la revendication 2,
dans lequel lesdits moyens de mappage sont adaptés pour appliquer une fonction de rotation pour faire tourner un premier desdits deux symboles entrants d'un angle de rotation tel qu'une première paire de symboles de canal

comprenne ladite version tournée dudit premier symbole entrant.

6. Codeur selon la revendication 2,
dans lequel lesdits moyens de mappage sont adaptés pour mapper bloc par bloc deux symboles entrant en deux paires de symboles de canal de sorte :

qu'une première paire de symboles de canal comprenne une version tournée de l'un des deux symboles entrants et une version cadrée de l'autre symbole entrant et
qu'une seconde paire de symboles de canal comprenne une version cadrée dudit un des deux symboles entrants et une version conjuguée logiquement inversée et complexe dudit autre symbole entrant.

7. Emetteur (10) pour transmettre des symboles de canal d'un courant de données de canal via un canal de transmission comprenant :

- un codeur selon la revendication 1 pour coder des symboles entrants d'un courant de données entrant en symboles de canal dudit courant de données de canal, et
- des moyens de transmission, en particulier deux antennes de transmission (11, 12), pour recevoir lesdits symboles de canal dudit codeur et transmettre lesdits symboles de canal via ledit canal de transmission.

8. Procédé de codage pour coder des symboles entrants d'un courant de données entrant en symboles de canal d'un courant de données de canal pour une transmission via un canal de transmission, le procédé de codage étant **caractérisé en ce qu'**il comprend les étapes consistant à :

- mapper bloc par bloc des symboles entrants en paires de symboles de canal, un bloc comprenant deux symboles entrants et étant mappé en deux paires de symboles de canal de sorte que lesdites deux paires de symboles de canal comprennent des versions cadrées desdits deux symboles entrants et/ou du conjugué complexe d'au moins l'un desdits deux symboles entrants, lesdites versions cadrées étant obtenues en appliquant une fonction de cadrage ayant un facteur de cadrage avec une valeur absolue différente de un et qui est linéaire pièce par pièce avec au moins deux pièces, et
- la délivrance desdits symboles de canal.

9. Décodeur adapté pour décoder bloc par bloc des symboles de canal reçus d'un courant de données de canal, qui ont été codés à partir de symboles entrants d'un courant de données entrant et transmis via un canal de transmission, dans lequel, au cours du codage, des symboles entrants ont été mappés bloc par bloc en paires de symboles de canal, un bloc comprenant deux symboles entrants et étant mappé en deux paires de symboles de canal de sorte que lesdites deux paires de symboles de canal comprennent des versions cadrées desdits deux symboles entrants et/ou du conjugué complexe d'au moins l'un desdits deux symboles entrants, lesdites versions cadrées étant obtenues en appliquant une fonction de cadrage ayant un facteur de cadrage avec une valeur absolue différente de un et étant linéaire pièce par pièce avec au moins deux pièces, ledit décodeur comprenant en outre :

- des moyens de sélection pour sélectionner une paire de valeurs de fonction possibles de symboles entrants pour décoder un bloc courant de symboles de canal reçus comprenant une paire de symboles de canal reçus y et
- des moyens de soustraction pour déterminer des premières versions cadrées intermédiaires $D_2(s)$ de ladite paire sélectionnée de valeurs de fonction possibles de symboles entrants s en appliquant une sous-fonction D_2 et pour soustraire des termes comprenant lesdites premières versions cadrées intermédiaires $D_2(s)$ de ladite paire sélectionnée de valeurs de fonction possibles de symboles entrants s à partir de ladite paire de symboles de canal reçus y pour obtenir des deuxièmes versions intermédiaires z de ladite paire sélectionnée de valeurs de fonction possibles de symboles entrants s,
- des moyens de détection pour détecter des troisièmes versions intermédiaires s de ladite paire sélectionnée de valeurs de fonction possibles de symboles entrants s en appliquant une détection de force zéro,
- des moyens de calcul pour calculer la distance euclidienne entre le signal reçu et les symboles estimés,
- des moyens de tranchage pour trancher lesdites troisièmes versions intermédiaires estimées s de ladite paire sélectionnée de valeurs de fonction possibles de symboles entrants s pour obtenir des estimations de ladite paire sélectionnée de valeurs de fonction possibles de symboles entrants s, et
- des moyens de commande pour répéter lesdites étapes avec d'autres paires de symboles entrants possibles s jusqu'à ce qu'un état d'arrêt prédéterminé soit rencontré ou jusqu'à ce qu'une distance euclidienne minimale soit trouvée et pour délivrer ladite paire de symboles entrants possibles s entraînant la distance euclidienne minimale.

10. Récepteur (20, 50) pour recevoir des symboles de canal d'un courant de données de canal via un canal de transmission comprenant :

- des moyens récepteurs, en particulier une ou deux antennes réceptrices (21, 51, 52), pour recevoir ledit courant de données de canal via ledit canal de transmission et pour délivrer des symboles de canal dudit signal de données de canal à un décodeur et
- un décodeur selon la revendication 9 pour décoder bloc par bloc les symboles de canal reçus dudit courant de données de canal.

11. Signal de données codées pour transmission via un canal de transmission acheminant des symboles de canal d'un courant de données de canal, qui ont été codés à partir de symboles entrants d'un courant de données entrant, dans lequel, au cours du codage, les symboles entrants ont été mappés bloc par bloc en paires de symboles de canal, un bloc comprenant deux symboles entrants et étant mappé en deux paires de symboles de canal de sorte que lesdites deux paires de symboles de canal comprennent des versions cadrées desdits deux symboles entrants et/ou du conjugué complexe d'au moins l'un desdits deux symboles entrants, lesdites versions cadrées étant obtenues par application d'une fonction de cadrage ayant un facteur de cadrage avec une valeur absolue différente de un et qui est linéaire pièce par pièce avec au moins deux pièces.

12. Produit de programme d'ordinateur comprenant des moyens de codage de programme pour amener un ordinateur à effectuer les étapes du procédé selon la revendication 8, lorsque ledit programme d'ordinateur est effectué sur un ordinateur.

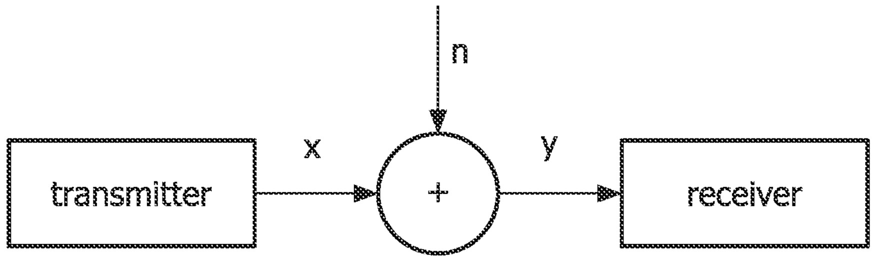

n
x
y
transmitter
+
receiver

FIG. 1

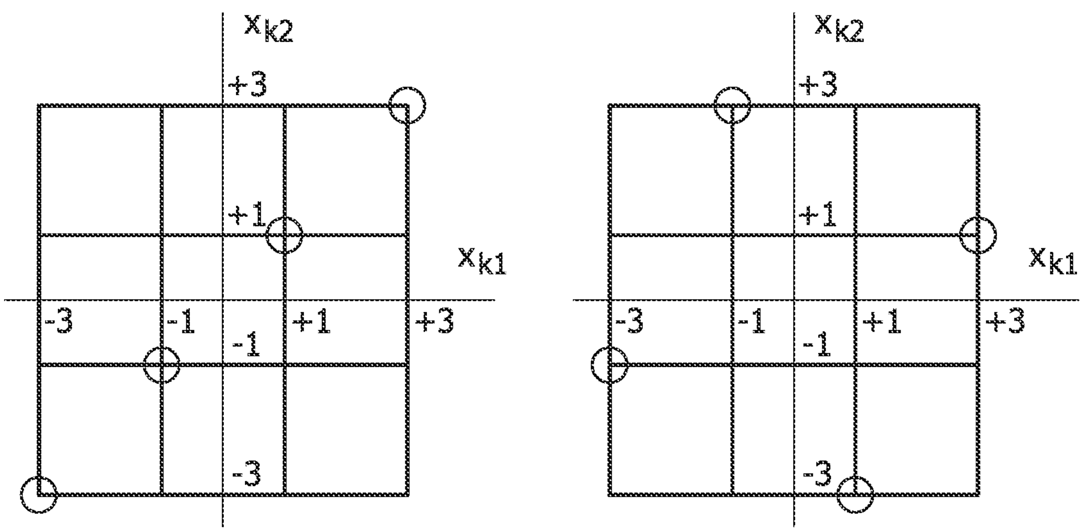

x_{k2}
+3
+1
x_{k1}
-3
-1
+1
+3
-1
-3
x_{k2}
+3
+1
x_{k1}
-3
-1
+1
+3
-1
-3

FIG. 2

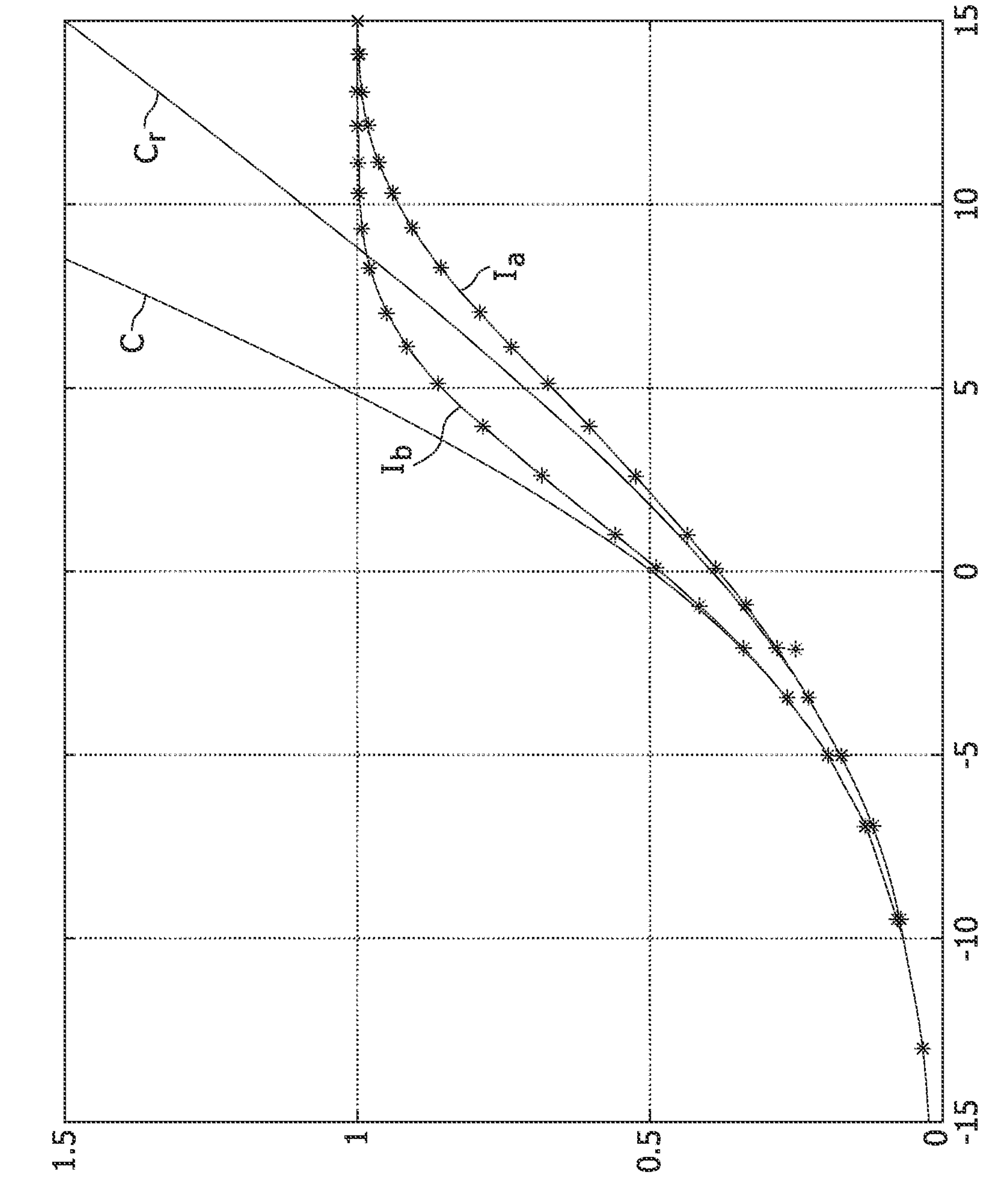


FIG. 3

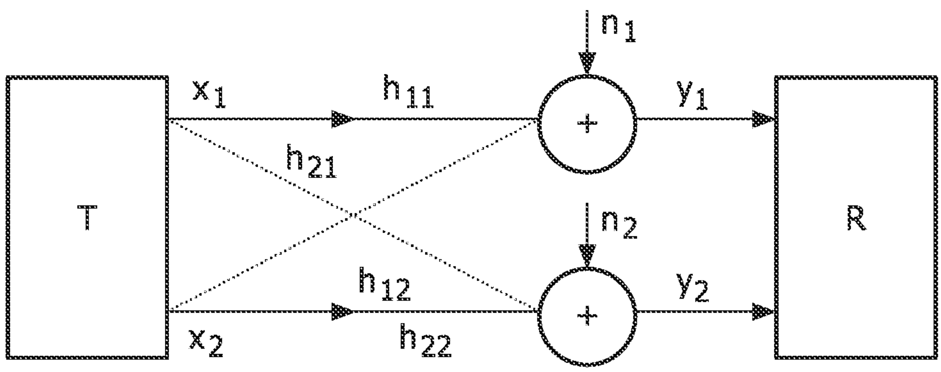


FIG. 4

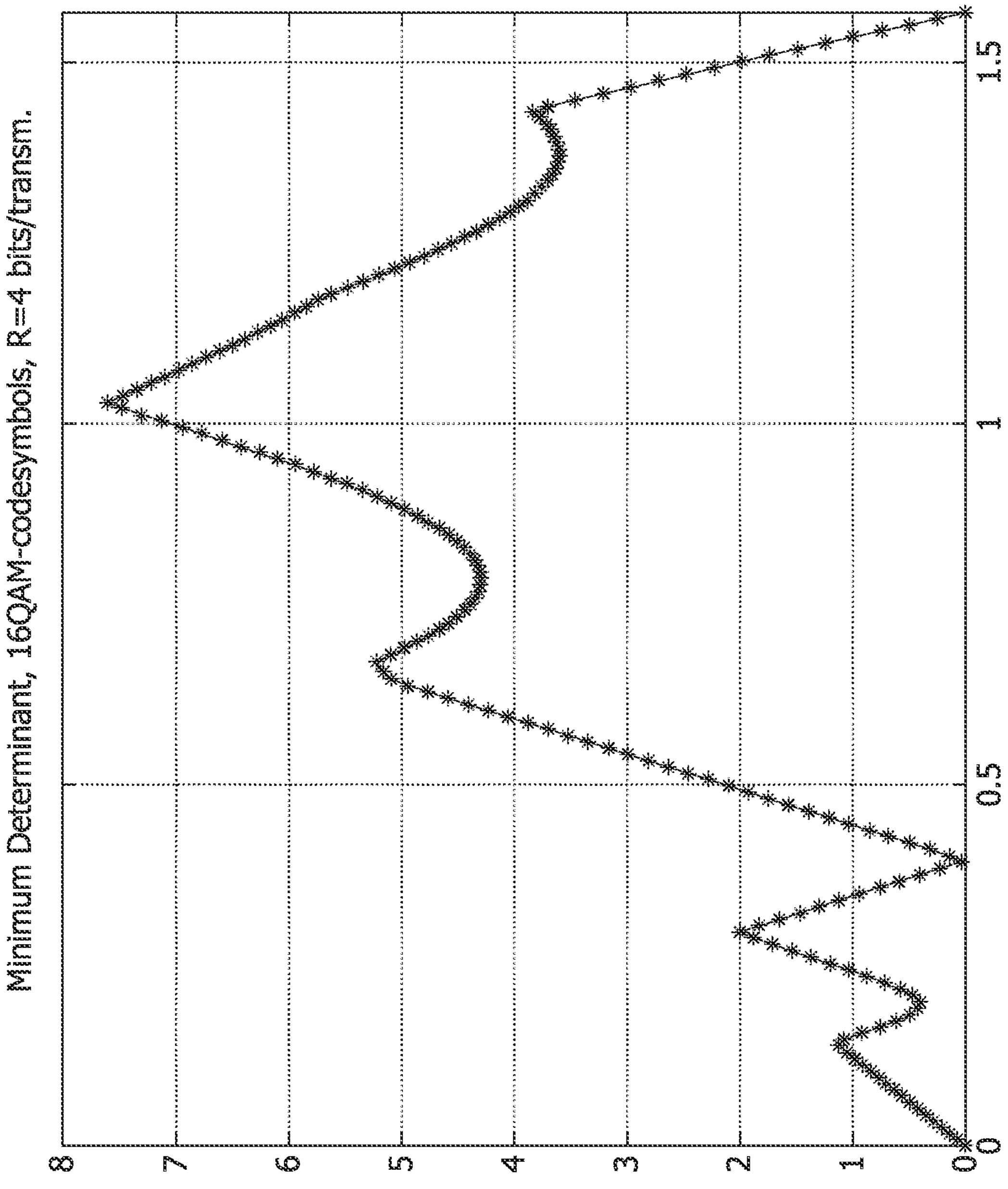

Minimum Determinant, 16QAM-codesymbols, R=4 bits/transm.
0
1
2
3
4
5
6
7
8
0
0.5
1
1.5

FIG. 5

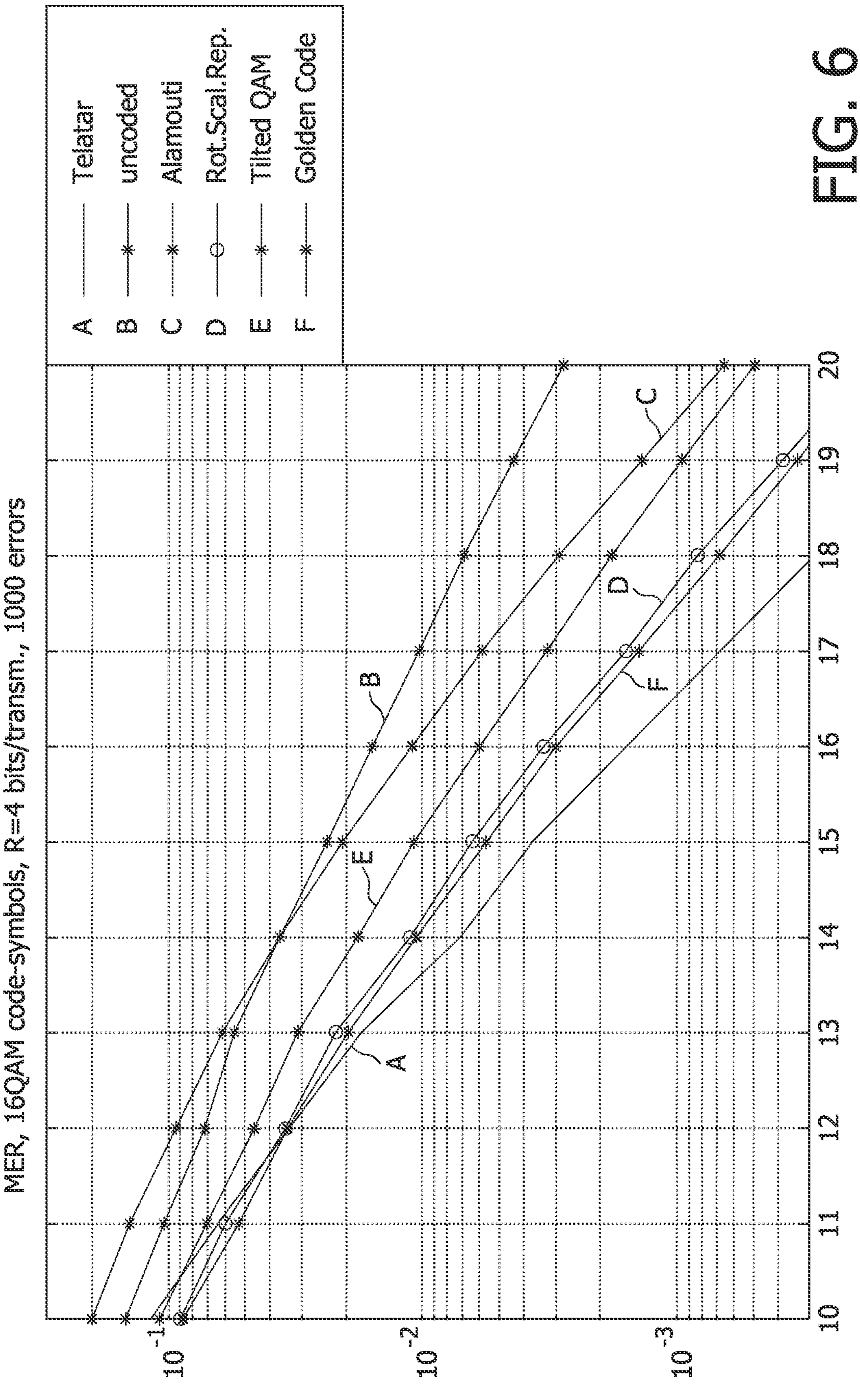

MER, 16QAM code-symbols, R=4 bits/transm., 1000 errors
A Telatar
B uncoded
C Alamouti
D Rot.Scal.Rep.
E Tilted QAM
F Golden Code
10^{-1}
10^{-2}
10^{-3}
10
11
12
13
14
15
16
17
18
19
20
A
B
C
D
E
F

FIG. 6

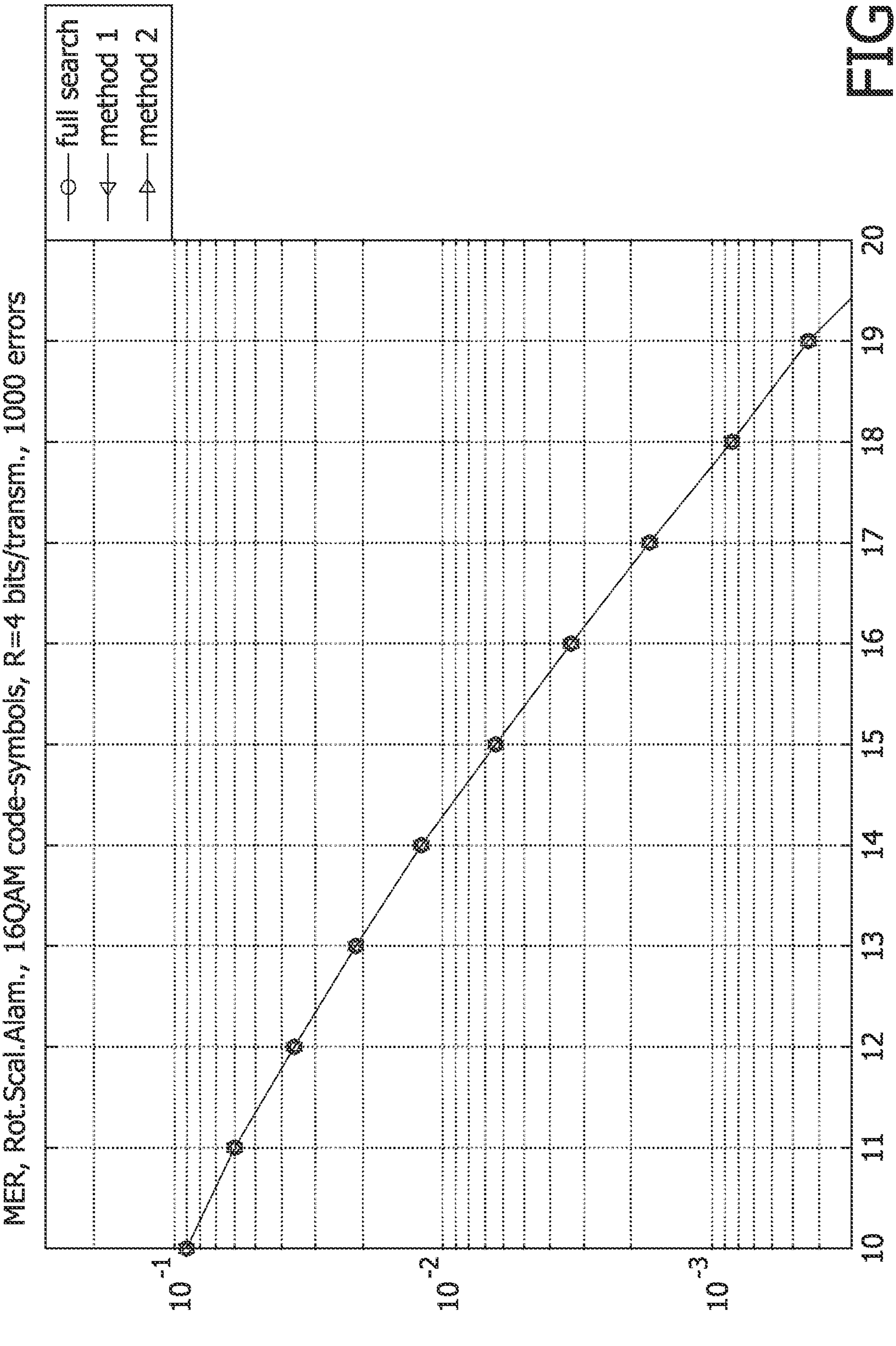
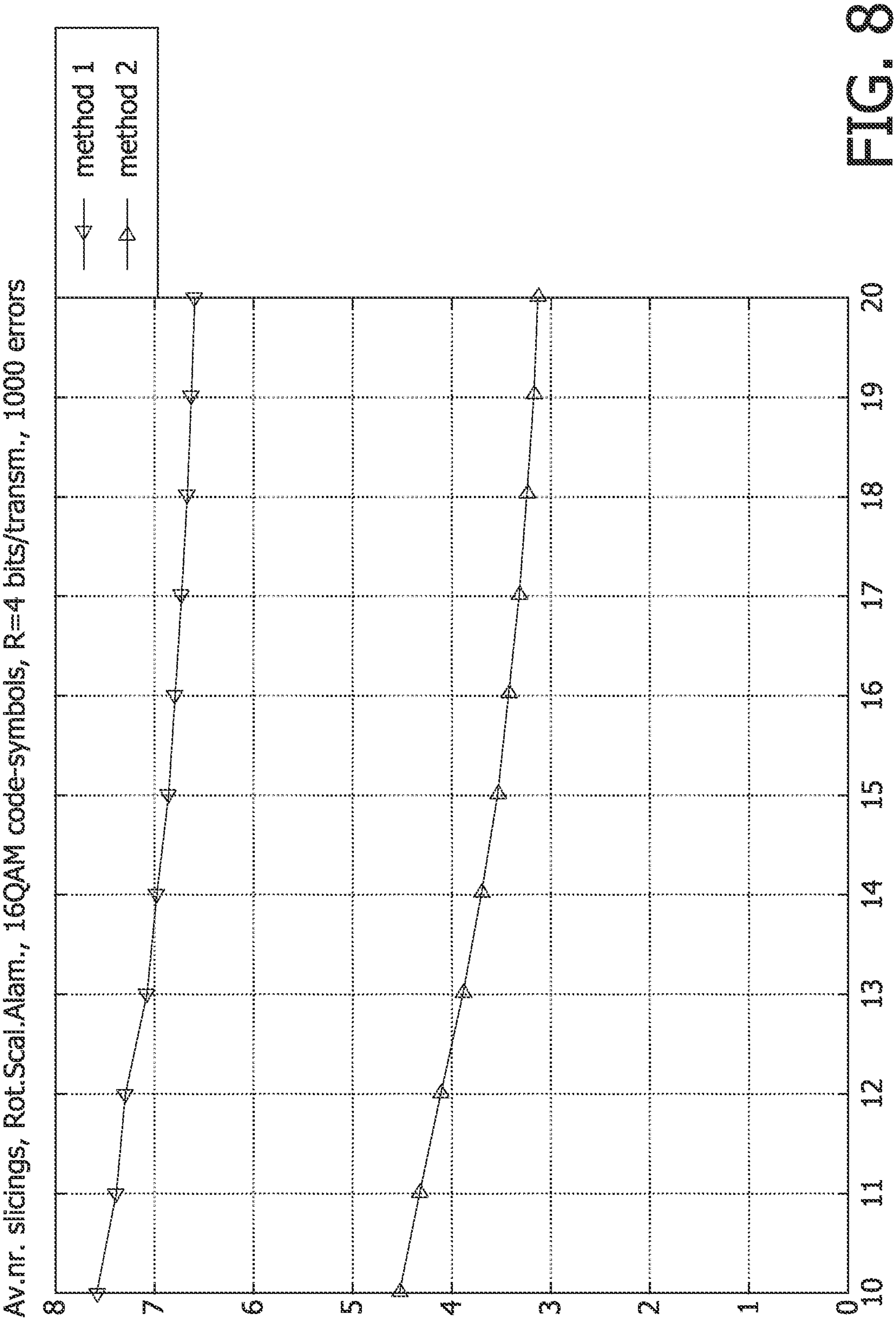

MER, Rot.Scal.Alam., 16QAM code-symbols, R=4 bits/transm., 1000 errors
full search
method 1
method 2
10^{-1}
10^{-2}
10^{-3}
10
11
12
13
14
15
16
17
18
19
20

FIG. 7


Av.nr. slicings, Rot.Scal.Alam., 16QAM code-symbols, R=4 bits/transm., 1000 errors
method 1
method 2
0
1
2
3
4
5
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7
8
10
11
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19
20

FIG. 8

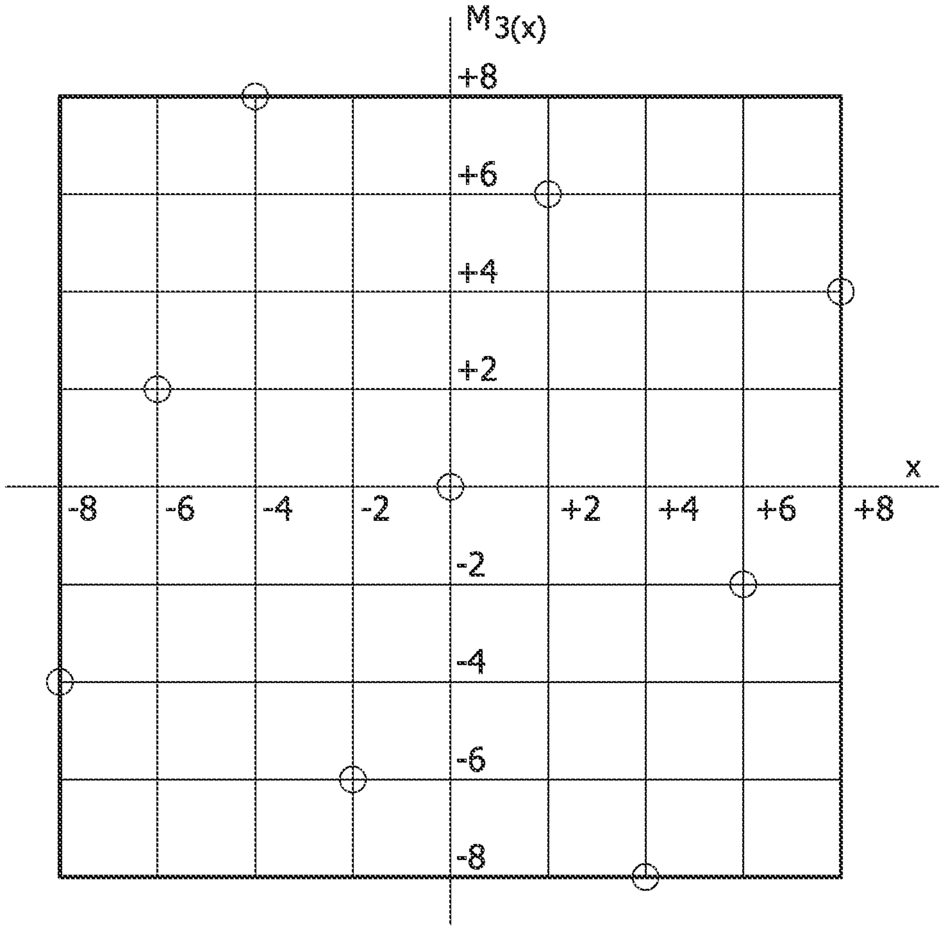

FIG. 9

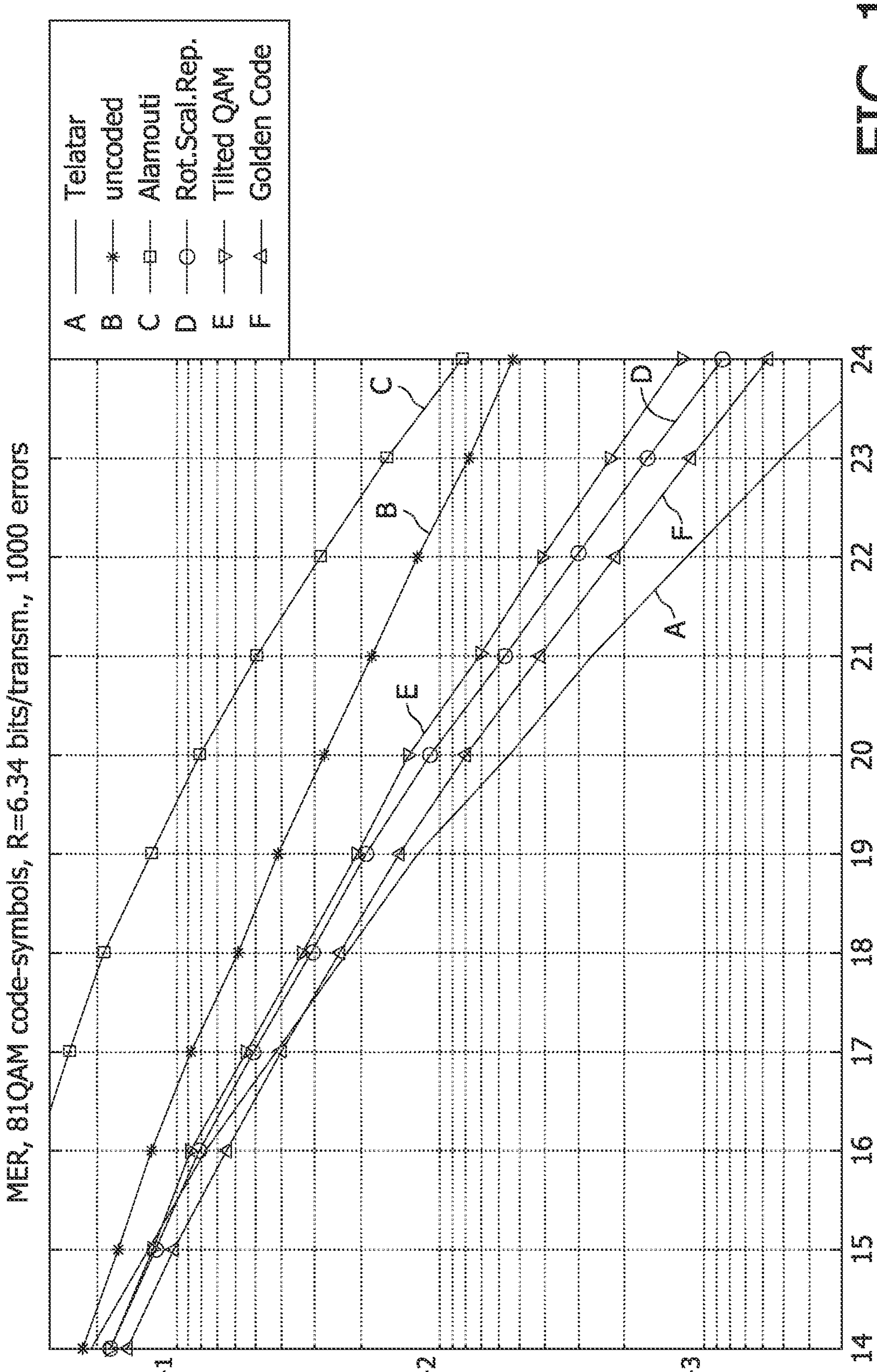
MER, 81QAM code-symbols, R=6.34 bits/transm., 1000 errors
A Telatar
B uncoded
C Alamouti
D Rot.Scal.Rep.
E Tilted QAM
F Golden Code
10^{-1}
10^{-2}
10^{-3}
14
15
16
17
18
19
20
21
22
23
24
A
B
C
D
E
F

FIG. 10

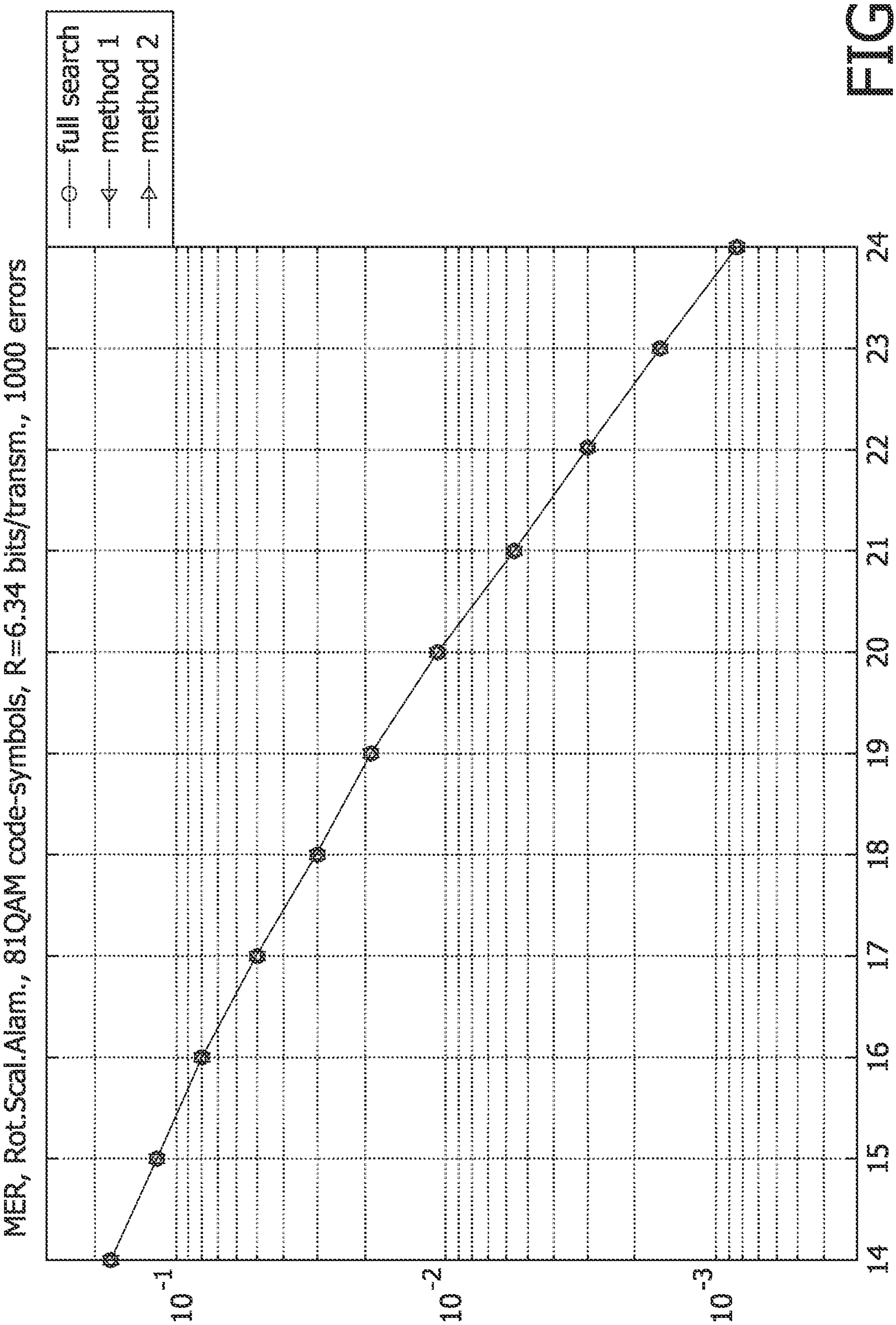
MER, Rot.Scal.Alam., 81QAM code-symbols, R=6.34 bits/transm., 1000 errors
full search
method 1
method 2
10^{-1}
10^{-2}
10^{-3}
14
15
16
17
18
19
20
21
22
23
24

FIG. 11

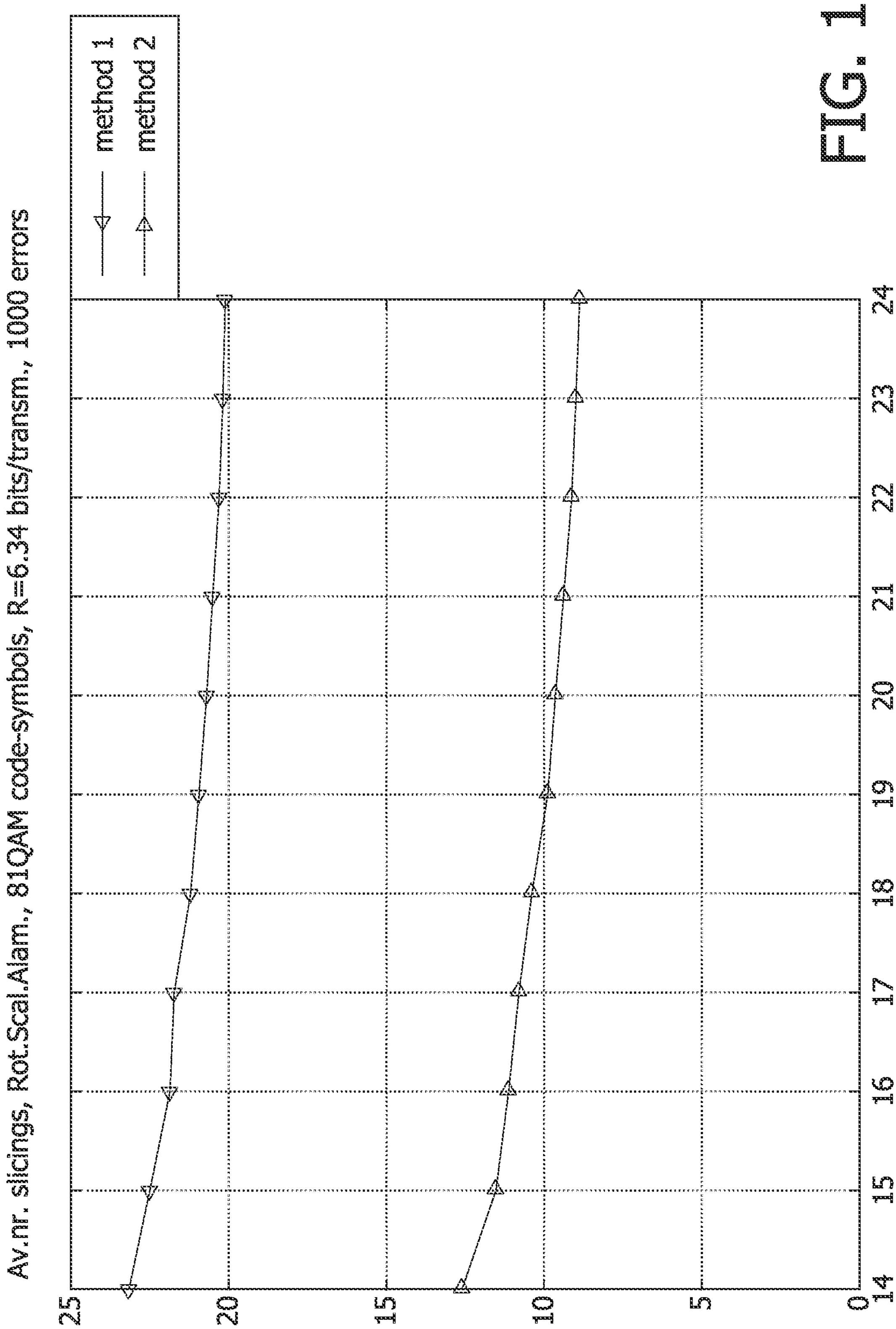
Av.nr. slicings, Rot.Scal.Alam., 81QAM code-symbols, R=6.34 bits/transm., 1000 errors
method 1
method 2
25
20
15
10
5
0
14
15
16
17
18
19
20
21
22
23
24

FIG. 12

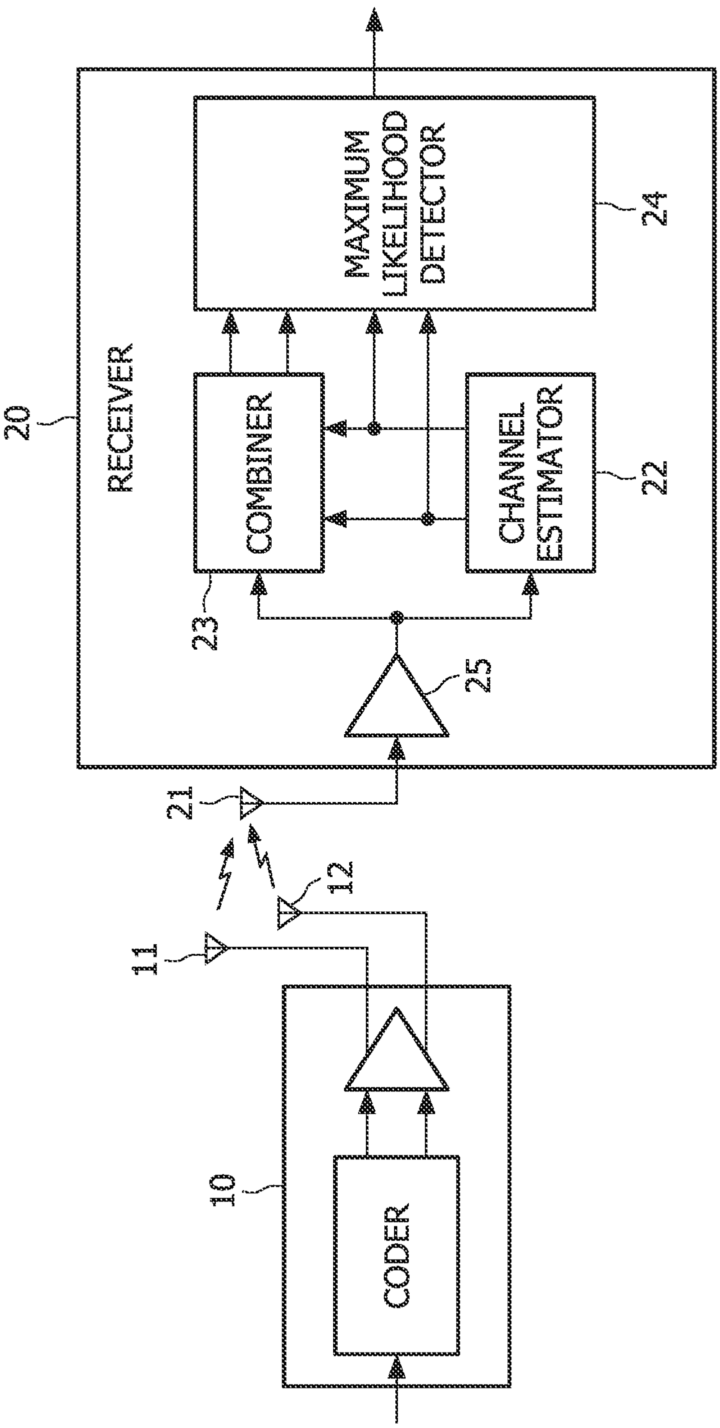
20
RECEIVER
MAXIMUM
LIKELIHOOD
DETECTOR
24
COMBINER
CHANNEL
ESTIMATOR
23
22
25
21
12
11
10
CODER

FIG. 13

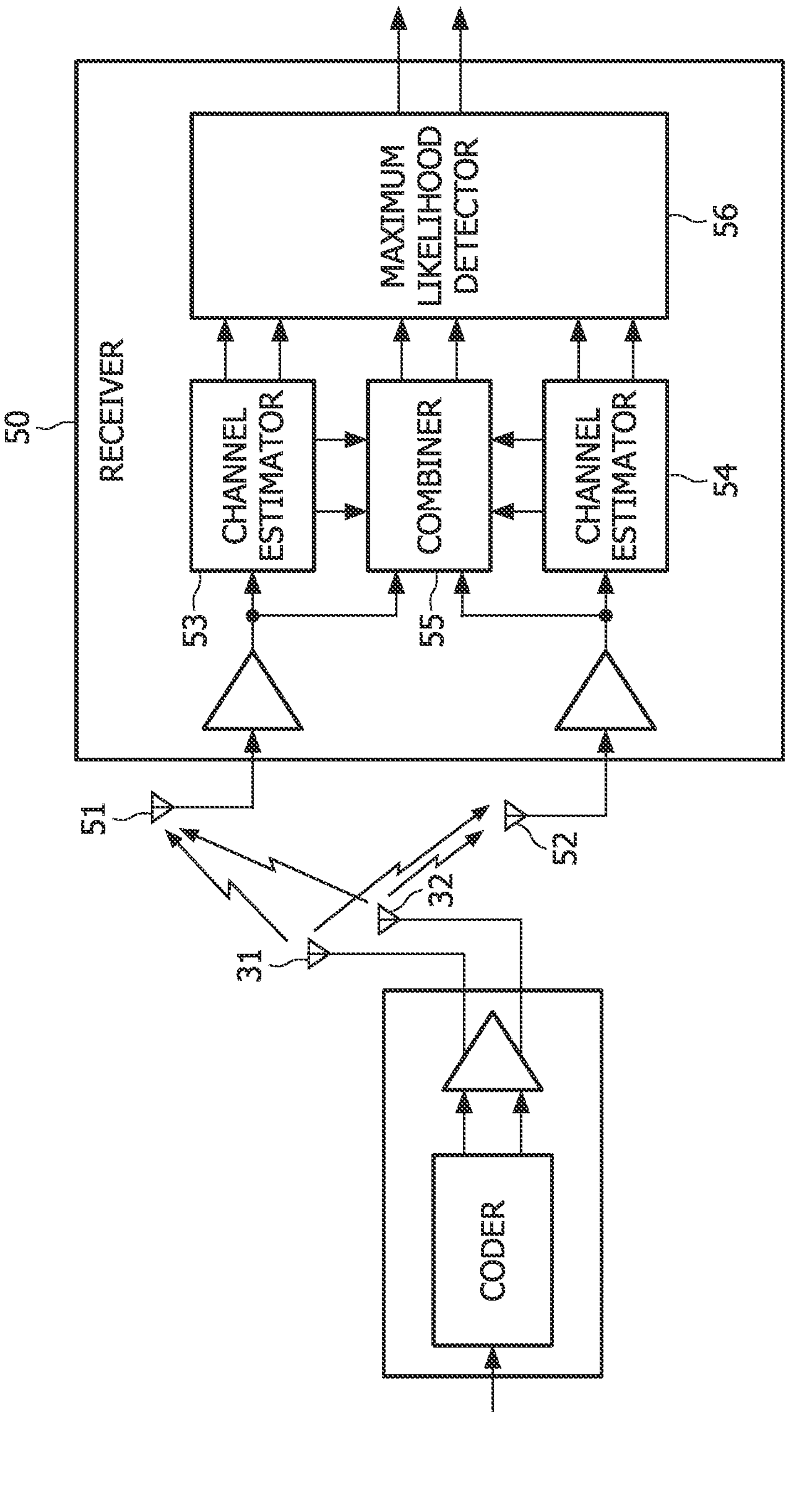
50
RECEIVER
CHANNEL ESTIMATOR
COMBINER
CHANNEL ESTIMATOR
MAXIMUM LIKELIHOOD DETECTOR
53
55
54
56
51
52
31
32
CODER

FIG. 14

REFERENCES CITED IN THE DESCRIPTION

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